D. M. NEALE

Cold Cathode Tube Circuit Design

CHAPMAN & HALL

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D. M. NEALE, B.Sc., A.M.I.E.E.
Head of Electronics Laboratory, Research Department,
Ilford Limited, Brentwood, Essex

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Preface

Whereas numerous excellent works are available already dealing with thermionic and semiconductor devices and circuit techniques, cold cathode devices have remained strangely neglected.

In many industrial applications, trigger and counting tubes offer advantages of circuit simplicity, economy or reliability in comparison with vacuum or solid-state devices. Although the choice of active element should in each case be a purely technical decision, a designer will naturally think most freely in terms of components and techniques with which he is familiar. It is hoped that in presenting information on a wide range of cold cathode tubes and their application, this book will help circuit designers to acquire familiarity with a class of device today being used to a rapidly increasing extent.

Whilst the presentation has been arranged in a logical order, an attempt has also been made to keep each chapter self-contained. A reader seeking information on a particular class of device should thus find, without tedious cross-reference, a consecutive account of its characteristics, mode of operation, circuit design procedure, worked examples and 'cut and dried' circuits for a variety of typical applications.

Repeated emphasis has been laid on the need for fully-toleranced designs making allowance for the worst possible combination of variables. In the past, many installations have proved disappointing either because insufficient attention was given to this aspect or because the tubes themselves drifted out of tolerance. Adequate information on life stability is now available for modern close-tolerance tubes and, if it is applied properly, highly reliable circuits can be designed. In even a simple circuit, however, the number of variables can accumulate so quickly that the designer is tempted to shirk the solution of a corresponding number of simultaneous equations—assuming he can set them out in the first place. It is to remove this temptation that the Design Procedures have been set out. Use of these Design Procedures in the Laboratory has proved that, without expert knowledge, a technician can quickly derive a soundly engineered design for any of the problems con-

sidered. The layout of Worked Examples follows those of the Design Procedures so that any doubts as to interpretation may readily be resolved.

D. M. Neale

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CHAPTER ONE

Evolution of the Cold Cathode Tube

About 1700, less than twenty years after Otto von Guericke had built the first electrical machine, Newton and Hawksbee produced electrical discharges in glass spheres evacuated to a low pressure. It was not until 1856, however, that Heinrich Geissler produced the first discharge tubes bearing his name. These tubes variously contained air, carbon dioxide, and hydrogen at pressures of one or two millimetres of mercury. When high voltage a.c. was applied to sealed-in electrodes at either end luminous discharges of great beauty were produced, the colour of which depended on the nature of the gas filling. Scientifically, Geissler tubes were of interest for the study of radiation spectra. For other applications they were of limited value because the gases then available led to severe sputtering of the electrode material. As a result, the envelope blackened near the electrodes and the gas pressure fell progressively until, after a rather short life, the discharge could no longer be established with a reasonable applied voltage.

Neon was not discovered until 1898. For ten years thereafter the world's supply of pure neon was contained in two small tubes inverted over mercury and standing on the mantelpiece of Sir William Ramsay, its discoverer. As soon as Georges Claude began to isolate substantial quantities of a helium-neon mixture in 1908, however, the possibilities of the neon discharge tube as a light source were explored. At the Paris Motor Exhibition of 1910 Claude himself exhibited two 38-ft neon tubes. As in modern advertising signs, the light produced by these high-voltage tubes came from the positive column, the luminous region starting a centimetre or so from the cathode and extending to the anode.

Meanwhile work was in hand to develop a neon lamp for domestic use. Electricity was expensive at that time, and so also were filament lamps of even the lowest wattage rating. For certain purposes there would thus be an attraction in a relatively robust lamp of very low current consumption. Professor H. E. Watson has related [1] how in 'sparking-out' the impurities from Claude's helium/neon mixture, he observed the ease with which neon sustained an electrical discharge. In 1910 he determined the cathode potential fall for pure neon with

aluminium electrodes as 200 V. This meant that a tube without a further potential fall due to a positive column could be made to operate on the 220 V domestic supply then common in England. As such a lamp would have no positive column, the light would necessarily come from the negative glow surrounding the cathode. As a light source, such a lamp is inefficient. The negative-glow neon lamp provides only about 0·7 lumens/W as compared with about 20 lumens/W from high-voltage tubes similar to Claude's and about 8 lumens/W from the metal-filament incandescent lamps already available in 1910 [2].

Despite its limitations, the 'glow-lamp' soon found favour in Germany, where 220-V domestic supplies were common. As a result of research work directed by Dr F. Schröter, the German firm of Pintsch A.G. produced the first commercial glow-lamps in 1918. Two years later Osram G.m.b.H. were making 500 lamps a day [1]. These were used mainly as night-lights or switchboard indicators for which the low power consumption (3–5 W) was welcome, or in mines, where their relative ruggedness was an additional attraction.

Claude had experienced some difficulty with contaminants in his neon-filled tubes. Of the noble gases, he is reported as saying [3]: 'Neon has a very high idea of its dignity. Although capable of remarkable effects when isolated, it absolutely refuses to do anything when in contact with common gases.' Sputtering had also been a problem for which a solution had to be found. Techniques devised by Claude and others for the construction of high-voltage tubes were not all applicable to the manufacture of glow-lamps in quantity, however. It was thus a considerable advance when, in 1917, Pintsch A.G. showed that, in very small quantities, diatomic gases were a useful addition to the gas-filling. Not only was the breakdown potential lowered, but sputtering was reduced dramatically. Shortly afterwards the Dutch firm of N. V. Philips entered the story, recommending a helium–neon mixture containing up to 5% argon as a filling with which sputtering was greatly reduced.

In France and the U.S.A., the use of 110-V domestic supplies made the glow-tube of little interest unless it could operate on this low voltage supply. From 1914 the General Electric Laboratories at Schenectady had been making experimental lamps with this object, but by 1920 a satisfactory lamp had still not been developed. For a while there was a loss of interest in the glow-lamp in the U.S.A.

Osram G.m.b.H. was more successful and Dr Skaupy showed in 1917 that the cathode fall could be reduced sufficiently by coating the cathode with barium azide. Another five years were required before

these activated cathodes became a commercial proposition. In the meanwhile 1921 saw the first Philips neon glow-lamps for 110-V supplies. They used cathodes of magnesium and beryllium alloys, later improved by the addition of potassium nitrate.

Stimulated by the successes of their European competitors, the Americans returned to the problem and finally T. E. Foulke evolved at Schenectady two new coating processes. One was a variation of Skaupy's barium azide process. The other involved the conversion of barium carbonate to the oxide. From this renewed activity there emerged in 1929 the miniature neon lamp, of which the world consumption is now some 70 millions per annum [1, 3]. One of these tubes can now be sold for less than a shilling, and over its 25,000-hour life it will consume less than two-pennyworth of electric power. The objective of a cheap and economical tube has certainly been attained.

Even now, research on the neon glow-lamp continues. Lemaigre-Voreaux [4] has discussed the advent of long-life high-intensity lamps and the possibility of others operating with an anode–cathode voltage of only 23 V.

In developing the early neon lamps many factors had to be reconciled. Cathodes and gas-fillings had to be found which not only provided operation at reasonably low voltages without excessive sputtering, but also were mutually compatible and compatible with one of the limited range of materials then available for lead-out wires which could be taken through the glass envelope in a gas-tight seal. By the late 1920s, however, solutions to all these problems were available. The scene was thus set for the development of tubes with additional control electrodes.

John Logie Baird used a neon glow-lamp as the modulated light source in his 1926 television receiver. As the modulation frequency did not exceed 10 kc/s, the response was sufficiently rapid, although the light output was necessarily extremely low.

Curiously, another early reference to control of a cold cathode discharge is also concerned with television. Manfred von Ardenne, writing in 1960 on the early days of television [5], mentions the development in 1925 of relaxation oscillators 'which could be synchronized by means of a cold cathode thyratron with external control from the signal'. Cobine [6] and others quote Knowles [7, 8] as the first to publish, in 1930, an account of a neon glow-tube with an anode–cathode discharge initiated by a third electrode. Knowles' tubes were essentially high-voltage devices, and so were of limited application.

The first cold cathode trigger tube to find extensive use was developed by Ingram at the Bell Telephone Laboratories. It was first announced

in 1936 [9] as 'the 313A Vacuum Tube' – despite the fact that it was filled with a mixture of neon and other gases to a pressure of several centimetres of mercury. This tube was of 'split-mushroom' construction: two identical half-hemispherical electrodes could be used interchangeably as trigger or cathode. The anode was a wire rod enclosed for most of its length in a glass sleeve. Both halves of the 'mushroom' were barium-activated and the tube could carry up to 30 mA at an anode–cathode voltage of 75 V. The 313A was developed for a specific application in telephony. In this it replaced a polarized relay and capacitor for selective ringing on party lines [10].

Only three years after the first trigger tubes became available Ingram was able to describe [11] most of the circuit techniques employed today. Since then advances have concerned tube design rather than circuitry. The 313A tube of 1936 had a rated life of only 300 hours of conduction. During the next decade improved manufacturing techniques led to an improvement in the life and stability of voltage stabilizing diodes and trigger tubes.

In 1946 Penning, Jurriaanse, and Moubis [12, 13, 14] described work done in the Philips Laboratories which led to a further remarkable improvement in both life and stability of the cold cathode tube. As a result, a new class of tube appeared having a cathode of pure molybdenum. During manufacture this is made to carry a discharge of such a value as to produce heavy sputtering. The surface layer of the cathode is thereby removed to reveal perfectly clean metal beneath. The sputtered-off material, moreover, cleans up any traces of contaminating gases in the filling, seals the surface of the glass envelope on which it settles, and serves as a 'getter' trapping any gas attempting to escape from the glass and liable to contaminate the cathode. The new technique was applied first to the 85A1 voltage reference diode, and during the next ten years to the production of high-stability trigger tubes [15].

In 1952 Hough and Ridler [16] described work which had been proceeding concurrently in the laboratories of Standard Telephones and Cables Ltd. Its prime object was to produce a trigger tube suitable for high-speed circuit applications. These demand a tube in which the gas-filling deionizes rapidly when the anode–cathode voltage is depressed. Additives to the gas-filling known to produce rapid deionization rapidly destroy an activated cathode, and so it was found necessary to return to pure metal cathodes – in this case, of nickel. At both the Philips and the S.T. & C. Laboratories it was also found desirable to include in the trigger tube envelope a small continuous discharge to provide the primary ionization needed for rapid and reliable triggering.

Elsewhere and at about the same time considerable effort was being directed towards further development of the activated-cathode tube. Here the objective was a mass-produced tube selling for a few pence apiece. Unlike the thermionic valve, a cold cathode tube does not deteriorate under stand-by conditions. It was therefore visualized that a cheap tube would be useful in computers and other circuits calling for large numbers of active elements. When an activated cathode carries current, however, the coated surface is progressively destroyed and the tube characteristics change accordingly. The cheap mass-produced tubes were no exception to this rule, and very few circuit designers made due allowance for ageing effects.

Inevitably the tubes concerned soon earned themselves a bad reputation. Undeservedly, the more sophisticated tubes using pure metal cathodes were to some extent stigmatized also. Thus the activated cathode, so valuable to the neon glow-lamp, rendered the trigger tube a disservice. A few trigger tubes using activated cathodes have proved highly successful, but in general the pure-metal type is now cheaper, longer-lived, and more stable. No doubt the next few years will lead to a balanced appraisal of their merits.

While the quest for the ideal cathode was in progress a novel type of cold cathode device was being developed in the S.T. & C. laboratories [16]. This had a number of similar cathodes, each of which accepted in turn the continuously flowing anode current, transfer from one cathode to the next occurring in response to pulses applied to intermediate electrodes. Such a multi-cathode stepping tube could combine in one envelope the functions of decade counter and display device. The first commercially available tube of this new class was, however, the Ericsson 'Dekatron', which appeared in 1950 [17]. Its success was immediate. For simplicity and reliability it was a big advance on circuits using thermionic tubes. Transistorized counters can now provide reliability superior to that of the earliest 'Dekatron' circuits, but, with the acquisition of experience and the introduction of new and improved tubes, stepping-tube circuit reliability has advanced to a comparable extent. In favourable applications the stepping tube has a working life of 100,000 hours. Based on a 40-hour week, this is longer than the probable working life of the circuit designer!

Over the last ten years an improved understanding of the physics of the gas discharge has brought a remarkable diversity of new tubes: stepping tubes operating up to 1 Mc/s; numerical indicators reminiscent of much earlier devices [18], but capable of displaying any digit from 0 to 9, substantially in the same position; stepping tubes capable of

switching these number tubes directly; other numerical indicators responding to signals as small as 5 V; tubes for switching high voltages; tubes for switching high currents; photographic flash and stroboscopic tubes – the list is continually lengthening.

It cannot yet be said that there is a tube for every purpose: it is neither reasonable nor sensible to hope there ever will be. But certainly there are many purposes for which the cold cathode gas discharge tube meets little or no competition. It is hoped that, with the assistance of the chapters which follow, the circuit designer will be able to familiarize himself with the various types available and so make full use of their unique properties.

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CHAPTER TWO

The Gas Discharge

For most practical purposes a gas in a weak electric field may be regarded as an insulator. In each atom the electrostatic charge of the nucleus is equal and opposite to that of the electrons surrounding it. The net charge on an atom or molecule is thus zero, each charge carrying with it an equal and opposite charge. Since a flow of current requires a transfer of *net* charge, no current can flow while this state persists.

There are, however, several processes by which a gas can be ionized, that is electrons can be removed from molecules or added to them. The resulting ions then carry net charges and, under the influence of electrostatic fields, they move. Thus current flows through a gas whenever an electrostatic field is applied across it *and* ions are present in the gas.

Of the various processes by which a gas may be ionized, not all are yet of practical value. For example, a carbon arc may be 'struck' by holding a lighted match under the gap between the carbons. Thermal agitation ionizes the gases in the flame, and these ions then move under the influence of the electrostatic field between the electrodes. Almost instantaneously, an arc discharge is initiated, i.e. a discharge having a current density at the cathode of the order of 100 A/cm^2.

The carbon arc was one of the first practical applications of electricity. Although it remains important today, its mechanism is still not fully understood. Much has been learnt about gaseous discharges at lower current densities, however, and a general theory of such discharges is now well established [1, 2].

When X-rays or gamma-rays pass through a gas a proportion of the photons is absorbed. Some of the energy is lost in expelling electrons from molecules of the gas, so producing electron–ion pairs. In the absence of an external electrostatic field electrons and ions tend to recombine. The chances of recombination clearly increase with the number of ions per unit volume. Equilibrium is thus reached when the number of free electrons is such that the rate of recombination becomes equal to the rate at which further electrons are being expelled.

If now a voltage is applied between two electrodes in the gas the electrostatic field attracts electrons towards the positive electrode and

ions towards the negative. Electrons reaching the positive electrode flow through the external circuit and replace other electrons which have neutralized positive ions striking the cathode. Thus an electron current flows through the gas and into the anode while an equal current of positive ions flows into the cathode. Measured at any point between the electrodes the current flowing through the gas is the same as that in the external circuit. But near the anode the current is predominantly due to electron flow while near the cathode it is almost entirely due to positive ion flow. This condition arises because, although electron–ion pairs are produced mainly between the electrodes, they immediately tend to move in opposite directions.

If the voltage between electrodes is low the acceleration of electrons and ions is correspondingly low. Consequently, many electrons are then recaptured by ions before they can reach the anode. As the inter-electrode voltage is increased, however, electrons and ions are accelerated more rapidly. Consequently, they reach the electrodes more quickly and the chances of recombination are reduced. This manifests itself as an increase in current with increase of inter-electrode voltage. It will be seen from Fig. 2.1 that the rise of current with voltage, though rapid at first, soon reaches a saturation value at which it remains practically constant, as shown by the part *AB* of the curve. It appears that over this range all the electrons liberated by the X-rays or gamma-rays reach the anode before any recombine with ions. The saturation current is thus a measure of the intensity of the ionizing radiation.*

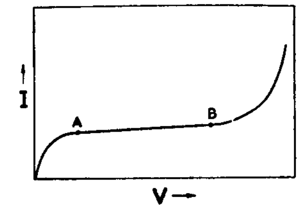


Fig. 2.1. Current–voltage characteristic of Townsend discharge.

Provided the rate of generation of free electrons or ions remains small, Fig. 2.1 is characteristic of the behaviour of any gaseous discharge. It is of no great importance how the charge carriers are injected. Thus, besides the mechanisms described already, photoelectric emission may

* This property is, in fact, used when alpha- or beta-radiation is measured with a gas-filled ionization chamber. Alpha and beta particles are readily stopped by gas molecules, X-rays and gamma-rays are not. An ionization chamber is therefore much less efficient as a detector of X- and gamma-rays. Both alpha and beta particles carry electric charges, so it is not surprising that their injection renders an otherwise non-conducting gas conducting. The ionization of a gas by X- and gamma-rays – however inefficiently – has here been described because these rays do not directly inject charge into the gas. They merely supply the energy needed to separate electrons from some of the molecules.

be responsible for the liberation of electrons. For clean, pure metal electrodes, photoelectric emission occurs only if the incident radiation is in the ultra-violet region. Composite electrodes, particularly those comprising monatomic layers of one metal on another, can provide photoelectric emission when illuminated by radiation in the visible part of the spectrum. In either event, photo-emission is of consequence only when it occurs at the cathode – an anode will immediately recapture its own electrons.

These factors are of importance in the operation of cold cathode tubes, since ions must be present in the gas before any discharge can occur. Usually there is sufficient ambient illumination to produce emission from a photosensitive cathode. From the few electrons so produced, a much larger discharge can result, as will shortly be described. When a tube is operated in darkness, however (as may happen inside an instrument case), the discharge may be delayed by seconds, or even minutes, until the gas is ionized by natural radiation such as a cosmic ray. If it is important to avoid such delays a small quantity of radioactive isotope is sometimes included in the envelope to ensure that ions are continually being produced from which a discharge can develop. Only low-energy isotopes are normally employed, and the emission is thus incapable of penetrating the envelope. With solid isotopes broken tubes might consitute a health hazard. For this reason tritium is a commonly used alternative.

Apart from deliberate illumination of the tube, there is one other well-established way of providing a continual supply of ions in the gas. This involves maintaining a small continuous discharge between the cathode and an auxiliary anode. The electrostatic field between these electrodes normally confines the ionization to their immediate vicinity. Other electrodes, such as the trigger electrode of a trigger tube, can be used to cause it to spread so that a discharge may be established between the cathode and other electrodes.

The Townsend Discharge

In the conditions considered so far the current through the gas is determined by the rate of injection of charge into the gas, either by alpha or beta particles or by ionization due to photons. This is known as the Townsend discharge. It is not self-maintained, and hence if the supply of ions ceases, the discharge stops, even though the inter-electrode voltage is maintained.

It was observed above that there is a range (from A to B in Fig. 2.1) over which the current is substantially independent of the inter-electrode voltage. If this saturation current represents the rate of injection of ions into the gas the increase of current beyond the point B must indicate the generation of further ions by the discharge itself. This occurs by two processes – gas multiplication and secondary emission. Each requires an inter-electrode voltage exceeding a value which depends on the pressure and nature of the gas filling. Secondary emission also depends on the material used for the cathode and the cleanliness of its surface.

An electron liberated in the gas is accelerated towards the anode by the positive potential thereon. The acceleration of the electron is proportional to the gradient of the electrostatic field acting on it, and accordingly the velocity of a freely moving electron may be expressed in terms of the potential through which it has been accelerated. In moving through a gas, however, an electron collides with gas molecules. So long as its momentum is below a certain level an electron bounces off a molecule like a dried pea bouncing off an egg: there is a change of direction, but very little loss of momentum. Thus, despite repeated collisions, electrons moving through a gas progressively acquire momentum on their way to the anode.

After an electron has been accelerated in this way through a potential known as the *ionizing potential* of the gas concerned it has acquired sufficient momentum to knock an electron from the outer electron shell of a molecule. It is then as if the dried pea hit the egg with sufficient energy to crack the shell. Momentum is lost in the collision because energy has been absorbed in changing the state of the shell. Thus immediately after the collision neither the original electron nor the newly ejected electron possesses enough momentum to produce further ionization. They are, however, accelerated by the electrostatic field so that, after again moving through the ionizing potential, they are each capable of ionizing another molecule.

If the inter-electrode voltage is sufficiently high this process may be repeated several times before the electrons reach the anode. Ionization of gas molecules by colliding electrons is known as *gas multiplication*. In the gas-filled photocell it is used to increase by a factor of about 10 the number of electrons reaching the anode.

Secondary Emission

When an electron is knocked out of a gas molecule, either by a high-energy photon or by a colliding electron, the atom which has lost an electron carries an excess of positive charge, i.e. it is a positive ion. Such an ion is then accelerated towards the cathode by the electrostatic field between the electrodes. As it carries an equal (but opposite) charge,

an ion is subject to the same accelerating force as an electron, though the force acts in the opposite direction. The mean free path of an ion is much shorter than that of an electron, however. Moreover, because an ion has much the same mass as a gas molecule, it loses much of its momentum on each collision. Before it reaches the cathode, therefore, an ion seldom acquires sufficient momentum to ionize any of the molecules with which it collides. Some of the ions striking the cathode will, however, knock electrons out of the cathode surface. This process of secondary emission by positive ion bombardment is very inefficient: perhaps 1% of the incident ions will liberate an electron. Nevertheless, it is believed to be the mechanism which enables a discharge to become self-maintaining.

From what has gone before, the general behaviour of the Townsend discharge should now become clear. No discharge can occur until electrons are liberated in the gas. With a moderate electrode potential all these electrons are collected at the anode. Over a certain range the current is then substantially independent of the applied voltage. As the voltage is raised, however, the current begins to rise – slowly at first, then with increasing rapidity. This is due to a combination of two effects: the ionization of gas molecules by colliding electrons and the ejection of further electrons from the cathode by the positive ions bombarding it.

Breakdown

As the voltage is increased, a point is eventually reached at which, for each electron leaving the cathode enough electrons (and hence ions) are liberated by collisions to result, on average, in the ejection of one more electron from the cathode by positive ion bombardment. This process will be recognized as one of positive feedback, and it exhibits the usual property of 'run-away' once the loop gain becomes unity.

When ion bombardment of the cathode is just able to provide enough electrons to make the discharge self-maintaining therefore, the current increases almost instantaneously by several orders of magnitude. This process is known as *breakdown*. Breakdown may occur at a current as small as 10^{-12} A. Immediately after breakdown the current at the same voltage may have increased to 10^{-6} A. The discharge now assumes a negative-resistance characteristic, and to control the voltage and current it is thus necessary to include a large series resistance in the circuit. This implies that beyond breakdown the discharge is current-controlled. Accordingly, its characteristic is best presented with current shown as the independent variable, as in Fig. 2.2. Presented in this way, the

characteristic clearly shows that the discharge may be controlled by voltage only as far as the point C at which breakdown occurs.

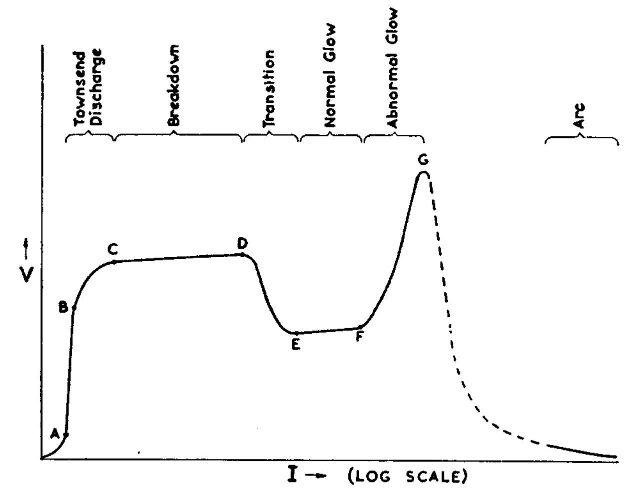


Fig. 2.2. Typical voltage–current characteristic of gas discharge.

Corona Discharge

If one electrode takes the form of a point or thin wire the high potential gradient surrounding this small electrode accounts for the greater part of the inter-electrode voltage. Gas multiplication in this region is then sufficiently effective to render the discharge self-maintaining. The positive-resistance characteristic of the greater part of the gap can still swamp the negative-resistance characteristic of the part surrounding the smaller electrode. Thus the discharge as a whole may still have positive-resistance characteristics, i.e. the part CD of the curve in Fig. 2.2 may still show a small positive slope. One class of tube – the corona stabilizer tube – has been developed to use this characteristic of the corona discharge for the stabilization of relatively high voltages (1 kV) at low currents (100 μA to 1 mA).

Effect of Space Charge

Until the point of breakdown is reached the current density in a discharge between moderately large, smooth electrodes is so low that there is negligible space charge. As shown by Curve 1 in Fig. 2.3, the potential gradient between electrodes is thus almost uniform, and in general this is less than the gradient giving maximum ionization efficiency.

Electrons and positive ions are liberated in a discharge at equal rates.

They are also removed at the same rate, once a stable condition has been attained. But on an average an electron reaches the anode in much less time than the positive ions take to reach the cathode. Consequently there is an excess of positive ions in the discharge. When breakdown occurs, the discharge current increases abruptly, and this space charge then becomes large enough to distort the electrostatic field appreciably. The gradient them becomes steeper near the cathode and less steep near the anode, as shown in Curve 2 of Fig. 2.3. It will be seen that, although the anode voltage remains the same, nearly all this voltage appears across less than half the electrode gap. At the distance from the cathode corresponding to the knee in Curve 2 there is a virtual anode. The potential gradient therefore increases over almost all the inter-electrode *voltage* (though over part only of the inter-electrode *distance*). This increased potential gradient allows electrons to acquire more energy between collisions with molecules. In this way more efficient ionization is obtained to sustain the increased current following breakdown.

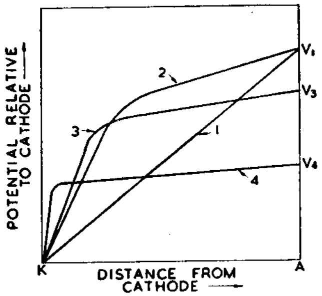


Fig. 2.3. Effect of space charge on potential distribution between cathode and anode.

If the external circuit resistance is now reduced so that the current increases, a further distortion of the potential distribution results. The virtual anode moves farther towards the cathode, as in Curve 3 of Fig. 2.3. A further increase in potential gradient ensues, leading to an ionization efficiency which has increased proportionately more than the current. As the current increases, the inter-electrode voltage therefore falls from V_1 to V_3: the discharge exhibits a negative resistance and is the transition stage indicated by *DE* in Fig. 2.2.

Further increase in discharge current still further reduces the inter-electrode voltage until a condition is reached (Curve 4 in Fig. 2.3) at which the potential gradient near the cathode is giving maximum ionization efficiency. At higher potential gradients both the number of collisions per centimetre and the probability of producing ionization on collision decrease as the potential gradient increases.

Normal Glow

A little consideration will show that the discharge will inherently seek the most efficient mode of operation. Any move towards inefficiency

will be deterred by that inefficiency; any process tending to increase the discharge current at a given voltage tends to be cumulative. As a consequence, the discharge naturally assumes a small cross-section so that the current density shall be that affording optimum ionization efficiency.

This is manifested in the nature of the glow which appears at the cathode. While the discharge remains in the transition stage indicated by *DE* in Fig. 2.2 this glow remains small. Once the condition has been reached providing maximum ionization efficiency, however, the cathode glow becomes progressively larger with increasing current. This is because the cross-sectional area of the current can thereby remain constant at the value corresponding to maximum ionization efficiency. A constant potential gradient implies a constant discharge voltage and, in fact, the inter-electrode potential rises only a few volts in the *normal glow* region (*E* to *F* in Fig. 2.2).

When the discharge current has increased so that the cathode glow occupies the whole of the cathode area any further increase in current necessarily increases the current density. This means the discharge can no longer operate under conditions favouring maximum ionization efficiency. Consequently, both potential gradient and inter-electrode potential rise as the current increases. The discharge then enters a region of *abnormal glow* (*F* to *G* in Fig. 2.2).

Light Output

The passage of a current through a gas causes electrons in some of the atoms to be pushed temporarily into new orbits. These new orbits represent higher energy levels. Consequently, as the electrons return to their normal orbits (ground states) energy is released, and this is apparent as light. The return from a higher energy level to the ground state may occur in several stages. Each transition from one energy level to another represents the release of a characteristic quantum of energy, and hence of light of a characteristic wavelength. As a result, the spectrum of a gas discharge reveals a number – sometimes a large number – of characteristic lines, the spacings and intensities of which are not uniform. The relative intensities of these lines are determined by the relative frequencies with which electrons make the transitions generating light of these particular wavelengths.

Although each gas has its own characteristic emission spectrum, a particular part of a discharge often emits light comprising only a proportion of the characteristic wavelengths. This is because, at a given distance from the cathode, electrons may not have acquired sufficient energy to excite the higher energy levels of atoms with which they collide.

Electrons leave the cathode surface with such a low velocity that a significant proportion recombine with ions. Just as energy is required to eject an electron from an atom, so recombination releases energy. Consequently, those electrons which recombine enter high energy-level orbits of the foster-parent ion. As they descend to lower energy levels, light is emitted. It is redder in colour than that from other parts of the discharge because the transitions between the higher energy levels of an atom are smaller – and therefore give rise to the emission of light of longer wavelengths – than the transitions between lower energy levels.

In other parts of the discharge electrons excite mainly only the lower energy levels. Before it can be accelerated sufficiently to excite a high energy level an electron usually suffers an inelastic collision with an atom, i.e. one in which it either ejects an electron or raises an electron above the ground state. In such a collision the colliding electron loses energy, and so must be accelerated once more before it can excite further atoms.

This process may occur repeatedly as an electron makes its way from cathode to anode. Under suitable conditions this is apparent as a series of striations in the glow of the discharge. The first of these, the negative glow, is close to the cathode – perhaps only 1 mm away. Beyond this is a relatively long dark space, the Faraday dark space, and then comes the positive column extending to the anode. Although the positive column is often strongly luminous and striated, in some tubes it may be only faintly luminous. In others the electrode configuration may cause it to be diffuse so that no striations are visible.

Arc Discharge

If the current in an abnormal glow discharge is progressively increased a point is reached at which the discharge becomes a self-sustained arc. This is characterized by an abrupt fall in inter-electrode potential to a few tens of volts and an increase in current to a density of 100–1,000 A/cm² at the cathode. The discharge no longer occupies the whole cathode surface, but concentrates on to a small part, the area of which is proportional to the current.

No exact theoretical treatment has yet been produced to explain all the features of the arc discharge. It does, however, appear reasonably certain that the high current density at the cathode can arise in at least two ways. If the cathode is of a refractory material, such as carbon, tungsten, or zirconium, thermionic emission from the cathode can yield current densities of the order of 100 A/cm². It is the bombardment of the cathode by positive ions which maintains the active area at a high temperature. Accordingly, such an arc stabilizes itself on to a single part of the cathode.

In low temperature arcs this is not the case. With a cathode of mercury or copper, for example, the discharge wanders ceaselessly over the cathode area. Such materials boil at temperatures below those at which thermionic emission could maintain the discharge. Here it seems the electrons are drawn from the surface of the cathode by the electric field. A sufficiently high electric field demands a current density of the order of 1,000 A/cm², and this is in fact observed.

Sputtering

It has been seen above that the effect of space charge leads to an increase in potential gradient at the cathode as the discharge current is increased. This gradient remains substantially constant while the discharge is in the normal glow region, but increases again as the abnormal glow is entered.

When the potential gradient is high, positive ions accelerated towards the cathode strike it with energy sufficient to drive off atoms of the cathode material. This sputtered metal is deposited on nearby surfaces and may cause trouble by contaminating adjacent electrodes or insulators.

Sputtering becomes more pronounced as the potential gradient is increased. It therefore tends to be serious if a heavy discharge current is passed. In practice, it is often necessary to guard against accidental reversal of electrode potential. An electrode unintentionally used as a cathode may require very heavy bombardment by positive ions in order to yield the secondary electrons needed to maintain a discharge. Such heavy bombardment will almost certainly lead to sputtering.

Gas molecules can be trapped on the surfaces on to which the sputtered material is deposited. In consequence, heavy sputtering produces a reduction in pressure of the gas-filling, an effect known as 'gas clean-up'. In early neon tubes gas clean-up was a serious problem. In modern trigger tubes using pure molybdenum cathodes, on the other hand, initial processing includes the deliberate production of heavy sputtering, as a result of which traces of contaminating gases are 'cleaned-up'.

REFERENCES

[1] PARKER, P. *Electronics*, E. Arnold, 1950.
[2] ACTON, J. R. and SWIFT, J. D. *Cold Cathode Discharge Tubes*, Heywood, 1963.

CHAPTER THREE

Diodes, Stabilizers, and Reference Tubes

Neon-filled cold cathode diodes are widely used as indicator and pilot lamps. Compared with tungsten filament lamps they have the following advantages:

(1) Current consumption may be a mere fraction of a milliampere. Neon lamps are thus suitable for indicating the state of h.t. supplies. They may also be controlled by small valves.

(2) The visible appearance does not depend critically on operating current. A change in current has no effect on the colour or brightness of the discharge: it affects only the area of the glow (and hence the total light output).

(3) The simplicity of the electrode structure lends itself to the production of a lamp more robust than a tungsten-filament lamp.

(4) Life can be very long, from 5,000 to over 50,000 hours and not shortened by repeated switching.

(5) Failure is usually a gradual process and is attended by a progressive blackening of the envelope. Replacement can thus be effected before complete failure occurs.

On the other hand, the following factors must be borne in mind:

(1) The brightness of the neon lamp is often relatively low.

(2) Reliable operation requires a supply in excess of 80 V.

(3) As the emission is almost completely confined to the yellow, orange, and red parts of the spectrum, a neon lamp can provide displays of these colours only; even if colour filters are used. Diodes filled with other gases can, of course, provide light of other colours. Indicators of this type are available from the Swiss company, Cerberus.

(4) After being switched off for some time the lamp may not strike immediately if it is in complete darkness. The delay varies from a few seconds to a few minutes.

Operation

The negative glow is the principal source of light from a neon lamp. When it is operated on d.c., therefore, only the negative electrode of

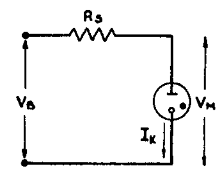
such a lamp will appear luminous. On an a.c. supply each electrode in turn serves as cathode on alternate half-cycles, and both electrodes accordingly appear luminous.

Indicators for operation on either a.c. or d.c. commonly have wire electrodes. Sometimes these comprise two parallel wires, sometimes they take the form of a rod and concentric ring. In either case the anode does not seriously obscure the cathode glow when the lamp operates on d.c. Lamps intended for a.c. operation may have a rod-and-cylinder construction. Although the cylinder may obscure the glow when the cathode is negative, the glow is clearly visible around the cylinder on the opposite half-cycle.

A current-limiting resistor must always be included in series with a neon indicator operating on d.c. This is because once a discharge is established, a neon lamp has a negative resistance characteristic (p. 12). With a.c. supplies a series choke or capacitor may be used, but a resistor remains the most usual choice.

Neons fitted with the domestic type of bayonet cap (B.C.) are com-

Fig. 3.1. Indicator diode and ballast resistance.



monly provided with an appropriate series resistor inside the cap. Neons with other types of cap are seldom so provided.

After striking, the current, I_K, of the lamp shown in Fig. 3.1 is given by:

$$I_K = \frac{V_B - V_M}{R_S} \quad (3.1)$$

where V_B = voltage of supply (d.c.).
V_M = maintaining voltage* of neon tube.
R_S = resistance in series with lamp.

From Equation (3.1) it will be seen that if the supply voltage does not greatly exceed the maintaining voltage a given change in the supply voltage produces a relatively large change in the tube current. On a d.c. supply this is seldom a problem, since, to ensure striking, the supply voltage should be at least $1\frac{1}{2}$ times the maintaining voltage.

* This is the inter-electrode voltage in the condition of normal glow (p. 14). It is also sometimes referred to as the 'burning voltage' or 'stabilizing voltage'.

With a.c. supplies it is necessary only that the peak voltage should reach this value. The average value of the tube current will then vary acutely with variations in supply voltage. This is evident from the curves marked I_1 and I_2 in Fig. 3.2 (*b*).

It is seldom necessary to calculate the average current in a neon tube

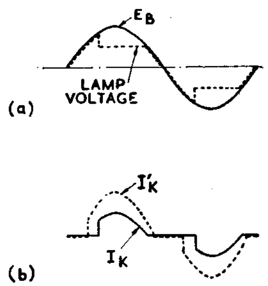


Fig. 3.2. Waveforms of diode operating on a.c. supply. (*a*) Supply voltage (E_B) and lamp voltage (V_M), and (*b*) corresponding current waveform, showing rapid increase from I_K to I_K' with increase in E_B.

on a.c. supply. When the figure serves any useful purpose a tedious calculation can be avoided by using Fig. 5.7 according to the procedure given in Chapter Eight with reference to character display tubes. More often the manufacturer's recommendations should be followed explicitly, since they are based on extensive practical tests.

Voltage Stabilizers

A gas discharge operating in either the corona or the normal glow mode provides an inter-electrode voltage almost independent of operating current. Voltage stabilizers (or regulators) are designed to exploit this characteristic so that a substantially constant voltage may be applied to a load, R_L in Fig. 3.3 (*a*), despite fluctuations in the supply voltage,

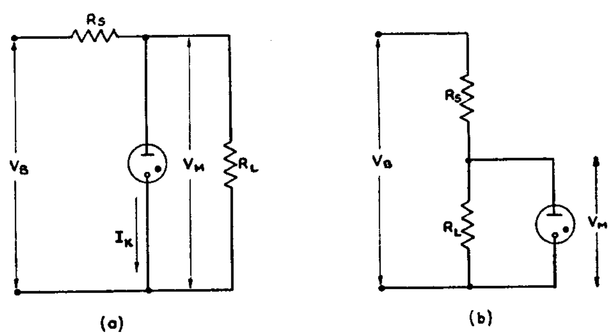


Fig. 3.3. Shunt stabilizer circuit (*a*) drawn conventionally, and (*b*) emphasizing problem of striking.

V_B. Alternatively, or additionally, the load R_L may vary, and the applied voltage will remain almost constant.

Corona Stabilizers

The construction and properties of tubes operating in the corona discharge mode have been described in some detail by Cohen and Jenkins [1, 2]. Such tubes comprise a cylindrical cathode with close-fitting glass or ceramic plugs at either end supporting a central wire anode. The whole assembly is enclosed in a glass envelope filled with spectroscopically pure hydrogen at a pressure not exceeding atmospheric. With given electrode dimensions the operating voltage of the tube is approximately proportional to the pressure of gas-filling. To avoid the need for pressure above atmospheric, however, high-voltage tubes are commonly made with larger cathode diameters. For each cathode diameter the anode wire diameter is so chosen that no close tolerance is required. Cohen and Jenkins describe a range of tubes covering the range 350 V to 7 kV with three electrode sizes: anode diameter 1 mm, cathode diameter 7 mm for 350 V to 2 kV; anode 1·2 mm, cathode 8 mm for 2 to 4·5 kV and anode 3 mm, cathode 13·2 mm for the range 4·5 to 7 kV.

Although it is not practicable to operate two or more corona stabilizers in parallel, the current rating of a tube may be increased by constructing it with an increased length of both anode and cathode. This makes a proportional reduction in the incremental resistance, R_I ($= dV_M/dI$).

The current range of a corona stabilizer is limited at the upper end by the danger of breakdown on to the glow discharge mode. In a high-voltage tube this is apparent as a pronounced fall in V_M. In low-voltage tubes the transition is much less obvious and, in fact, relaxation oscillations usually occur between the corona and glow states. These may be recognized with the aid of an oscilloscope which will reveal a saw-tooth voltage waveform appearing across the tube. Transition from corona to glow discharge occurs at a lower current when the electrodes are heated by extended operation at high current. The manufacturer's recommendations regarding maximum current should therefore be observed. Continued operation at high currents also increases the rate of gas 'clean-up' which reduces the gas pressure and hence V_M. As a result, the tube passes more rapidly out of tolerance. With moderate currents, the tube life is good. Cohen and Jenkins quote test data for 2-kV tubes suggesting a change in V_M of $2\frac{1}{4}\%$ over 10,000 hours operation at 300 μA.

The minimum tube current is usually of the order of a few micro-amperes. Below this current the discharge ceases to be self-maintaining. It extinguishes repeatedly, restriking each time the electrode potential and primary ionization rise sufficiently to initiate a further discharge 'avalanche'.

A corona stabilizer normally shows an incremental resistance, R_I, in the order of 100 kΩ. For a given tube, this varies with current and with method of measurement. R_I is greatest for rapid fluctuations of voltage or current, e.g. 50 c/s or more, which are too rapid to be followed by thermal effects in the tube. If a step-function is applied in a sense to increase tube current, however, the initial rise in V_M is decreased within a fraction of a second due to local temperature increase (and hence reduction in density) of the gas in the discharge. If R_I is calculated from readings taken about one second after an abrupt change, therefore, a lower value of R_I will result. From measurements made over a still longer period a still lower value of R_I is obtained, due to a further reduction in gas density in the discharge as the heated gas escapes to the space between cathode cylinder and outer envelope. This last value of R_I can even assume a negative value.

In general, R_I decreases with increase of tube current.

Glow-discharge Tubes

Voltage stabilizers operating in the mode of normal glow commonly provide anode–cathode maintaining voltages, V_M, in the range 60–180 V. Such tubes have cylindrical or planar cathodes and relatively small anodes in the form of wire rods. The value of V_M depends on the electrode geometry and the nature and pressure of the gas filling. Neon, argon, and neon–argon mixtures are all used as fillings in commercially available tubes. The cathode surface also plays an important part, an activated cathode providing a lower value of maintaining potential than one of pure metal.

The working range of current is limited at the lower end by instability of the discharge and at the other by the onset of increased sputtering as the discharge enters the region of abnormal glow. Within these limits a typical tube shows an incremental resistance, R_I $(= \mathrm{d}V_M/\mathrm{d}I)$ of a few hundred ohms. Voltage stabilizer tubes (as opposed to voltage reference tubes) are usually designed for operation over relatively large current ranges and to provide low values of R_I.

Stabilizers may be connected in series to provide voltages not obtainable from single tubes. In the G.E.C. 'Stabilovolt' tubes, several cup-shaped electrodes are mounted concentrically in a common envelope. The tube operates with a discharge between each adjacent pair of electrodes. The overall maintaining voltage of 280 V, which is obtained between the central anode and outer cathode, is thus the sum of four 70-V discharges in series. As five external connexions are available, stabilized outputs may be taken also from the intermediate 70-V, 140-V, and 210-V electrodes.

Benson [3], who has studied glow-discharge tubes over many years, has produced a valuable survey paper including a list of 272 references.

Striking

With a corona discharge tube, a self-maintaining discharge will develop with an applied voltage only a few per cent in excess of the anode–cathode maintaining voltage, V_M, of the corona mode. For a tube to operate in the normal glow mode, however, a voltage must be applied sufficient to take the discharge through the corona region and to the point of breakdown. The current then increases abruptly and the tube 'strikes'.

With a simple glow-discharge diode up to $1\frac{1}{2}$ times the maintaining voltage must be applied to produce striking. In the dark a still higher voltage may be required unless a radioactive substance is included in the tube. A trace of tritium is sometimes included in the gas filling for this purpose.

Circuit arrangements which are satisfactory in other respects often fail to provide striking of the tube on load. This is due to the load and series resistor forming a potentiometer reducing the available voltage below that needed for striking. When Fig. 3.3 (*a*) is redrawn as Fig. 3.3 (*b*), the effect is obvious.

Striking will occur if V_B is at least $(R_S + R_L)/R_L$ times the striking voltage, V_{IG}. Hence,

$$V_B/V_{IG} \geqslant 1 + (R_S/R_L) \qquad (3.2)$$

Relation (3.2) must be satisfied for the lowest value of R_L for which striking is required.

For a given value of R_L, the minimum value of R_S is given by,

$$V_{IG(max)} - V_{M(min)} = [I_{K(max)} - (I_{L(max)} - I_{L(min)})] \cdot \frac{R_{S(min)} \cdot R_{L(min)}}{R_{S(min)} + R_{L(min)}}$$

whence

$$\frac{R_{L(min)}}{R_{S(min)}} = \frac{I_{K(max)}R_{L(min)} - [I_{L(max)} - I_{L(min)}]R_{L(min)}}{V_{IG(max)} - V_{M(min)}} - 1$$

Substituting in Relation (3.2), and putting

$$R_{L(min)} = \frac{V_{M(min)}}{I_{L(max)}}$$

$$\frac{V_{B(min)}}{V_{IG(max)}} \geqslant \frac{I_{K(max)} + I_{L(min)} - I_{L(max)}}{I_{K(max)} + I_{L(min)} - q \,.\, I_{L(max)}} \tag{3.3}$$

where $q = V_{IG(max)}/V_{M(min)}$.

The value of q is typically 1·2–1·3, but it exceeds 1·5 with some tubes.

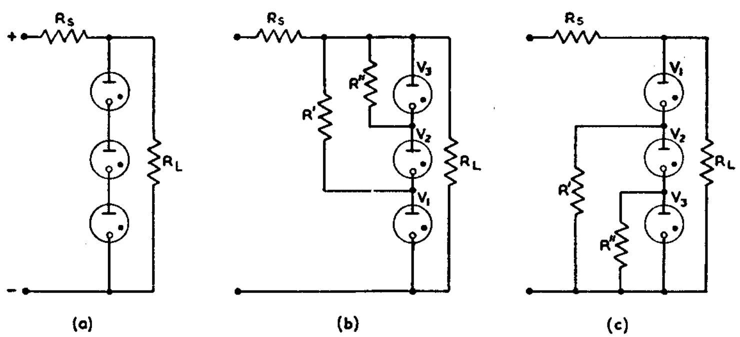


Fig. 3.4. Series-connected stabilizers (*a*) without, (*b*) and (*c*) with, resistances assisting striking.

When several similar diodes are connected in series, as in Fig. 3.4 (*a*), qV_M must be applied to each tube simultaneously to ensure striking. When n diodes are used, therefore, the voltage across the load must momentarily rise to qnV_M. This is not always either convenient or desirable. By connecting relatively high resistances R', R'' (usually $\frac{1}{2}$–1 MΩ) from each diode junction to either stabilized rail, as in Figs. 3.4 (*b*) and (*c*), the necessary excess voltage is usefully reduced without materially affecting performance subsequent to striking. Initially R' shunts V_2 and V_3 so that V_1 strikes. R'' now shunts V_3 so that V_2 strikes also. With V_1 and V_2 already struck, V_3 will strike when the output voltage has risen to only $(n - 1 + q)V_M$ instead of the qnV_M required by the arrangement in Fig. 3.4 (*a*).

Voltage stabilizers are available containing a third electrode used as a priming anode. Such tubes make it easier to ensure reliable striking because the full supply voltage – or an alternative (even higher) supply – may be applied to the priming anode through a current-limiting resistor, R_P (commonly 0·1–0·3 MΩ), in Fig. 3.5. The resulting discharge between priming anode and cathode ensures that the main anode-to-cathode discharge will strike at a voltage only about 10% above the

maintaining voltage, V_M. Even when reliable striking is not in itself a problem, such tubes can be useful because they lead to only a small excess of output voltage before striking occurs.

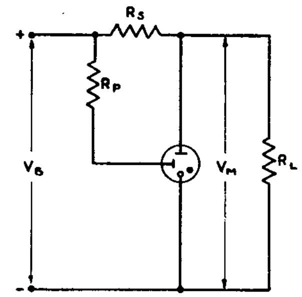


Fig. 3.5. Primed stabilizing diode.

Operating Current and Voltage Range

In general, a load, (R_L), is connected in parallel with a stabilizer, as in Fig. 3.3. It can then be seen on inspection that the current, I_K, drawn by the tube is given by:

$$I_K = \frac{V_B - V_M}{R_S} - \frac{V_M}{R_L} \tag{3.1a}$$

It will be noted that Equation (3.1a) is the more general form of Equation (1), to which it reduces when $R_L \gg R_S$.

Writing I_L for V_M/R_L, Equation (3.1a) provides:

$$I_K + I_L = \frac{V_B - V_M}{R_S} \tag{3.4}$$

This means that, for constant values of V_B, V_M, and R_S, the tube current varies in a complementary manner to the load current. The working current range of a stabilizer must therefore exceed the anticipated load current variations.

Full design of a stabilizing stage requires consideration of the effects of tolerances in V_B, V_M, and R_S. The 'worst × worst' case must be considered. Thus:

$$I_{K(max)} + I_{L(min)} \geqslant \frac{V_{B(max)} - V_{M(min)}}{R_{S(min)}} \tag{3.4a}$$

and

$$I_{K(min)} + I_{L(max)} \leqslant \frac{V_{B(min)} - V_{M(max)}}{R_{S(max)}} \tag{3.4b}$$

The manufacturer's information on a given tube type provides limiting values of I_K and V_M. Usually the range of I_L is known also and the *percentage* swing in V_B and the percentage tolerance of R_S. Thus Relations (3.4a) and (3.4b) yield simultaneous equations from which minimum nominal values of V_B and R_S may be determined.

If t_1 is the fractional positive tolerance on the nominal value of R_S
t_2 ,, ,, negative ,, ,, ,, R_S
r_1 ,, ,, positive ,, ,, ,, V_B
r_2 ,, ,, negative ,, ,, ,, V_B
s_1 ,, ,, positive ,, ,, ,, V_M
s_2 ,, ,, negative ,, ,, ,, V_M

then Relations (3.4a) and (3.4b) may be used to show that

$$V_B \geqslant \frac{\{(1+s_1)(1-t_2)[I_{K(max)}+I_{L(min)}] - (1-s_2)(1+t_1)[I_{K(min)}+I_{L(max)}]\}V_M}{(1-r_2)(1-t_2)[I_{K(max)}+I_{L(min)}] - (1+r_1)(1+t_1)[I_{K(min}+I_{L(max)}]} \qquad (3.5)$$

and

$$R_S \geqslant \frac{[(1+r_1)(1+s_1)-(1-r_2)(1-s_2)]V_M}{(1-r_2)(1-t_2)[I_{K(max)}+I_{L(min)}] - (1+r_1)(1+t_1)[I_{K(min)}+I_{L(max)}]} \qquad (3.6)$$

By putting $t_1 = t_2 = t$, Relation (3.5) may be re-written to express the resistor tolerance, t, in terms of the other parameters.

Thus
$$t \leqslant \frac{\left(\dfrac{V_B}{V_M}-1\right) - \left(\dfrac{r\,.\,V_B}{V_M}+s\right).\,n}{\left(\dfrac{V_B}{V_M}-1\right).\,n - \left(\dfrac{r\,.\,V_B}{V_M}+s\right)} \qquad (3.7)$$

where
$$r = r_1 = r_2$$
$$s = s_1 = s_2$$
and
$$n = \frac{[I_{K(max)}+I_{K(min)}]+[I_{L(max)}+I_{L(min)}]}{[I_{K(max)}-I_{K(min)}]-[I_{L(max)}-I_{L(min)}]}$$

For any value of V_B chosen in accordance with Relation (3.3), it is thus possible to calculate the corresponding value of t. If the value of t given by Relation (3.7) is negative the value of V_B must be increased until a satisfactory positive value is obtained.

Design Procedure for Diode Shunt Stabilizers

(Applicable also to Character Display Tube on D.C. Supply)

Reduced to a set of rules, the design procedure for a 'worst × worst' case is not so onerous as the foregoing might suggest. The procedure is as follows.

(*a*) Set out the following data:

Nominal tube maintaining voltage, V_M
Maximum tube ignition voltage, V_{IG}
(For a corona stabilizer, put ($V_{IG} = V_M$))
Minimum and maximum tube currents, $I_{K(min)}$ and $I_{K(max)}$
Minimum and maximum load currents, $I_{L(min)}$ and $I_{L(max)}$
Fractional tolerances, $+r_1$, $-r_2$, on the supply voltage, V_B
Fractional tolerances, $+s_1$, $-s_2$, on the maintaining voltage, V_M

(*b*) Evaluate

$$q = V_{IG}/(1-s_2)V_M$$

and
$$n = \frac{[I_{K(max)}+I_{K(min)}]+[I_{L(max)}+I_{L(min)}]}{[I_{K(max)}-I_{K(min)}]-[I_{L(max)}-I_{L(min)}]}$$

(*c*) Substitute values in Relation (3.3a) to determine the minimum value of V_B ensuring full-load striking:

$$V_B \geqslant \frac{V_{IG}}{(1-r_2)} \cdot \frac{I_{K(max)}+I_{L(min)}-I_{L(max)}}{I_{K(max)}+I_{L(min)}-q\,.\,I_{L(max)}} \qquad (3.3a)$$

(*d*) Select a convenient value of V_B in accordance with (*c*) above.

(*e*) Substitute in Relation (3.7) to determine the greatest permissible resistor tolerance, t.

(If $r_1 \neq r_2$, a mean value may be taken for r if the two values are not greatly different. Alternatively, a factor of safety may be added by adopting the larger value.)

$$t \leqslant \frac{\left(\dfrac{V_B}{V_M}-1\right) - \left(\dfrac{r\,.\,V_B}{V_M}+s\right).\,n}{\left(\dfrac{V_B}{V_M}-1\right).\,n - \left(\dfrac{r\,.\,V_B}{V_M}+s\right)} \qquad (3.7)$$

(*f*) If the value of t so obtained is negative or inconveniently small, repeat operations (*d*) and (*e*) above for a higher value of V_B.

(*g*) When a satisfactory (positive) value of t is obtained, calculate from Equation (3.4c) the corresponding nominal value of R_S.

$$R_S = \frac{2\left\{\left(1+\dfrac{r_1-r_2}{2}\right).\,V_B - V_M\right\}}{[I_{K(max)}+I_{K(min)}+I_{L(max)}+I_{L(min)}]} \qquad (3.4c)$$

(*h*) Select a standard resistance value of tolerance closer than t such that the extreme values will lie within the limits $R_S\,.\,(1 \pm t)$.

Note: If, for a given tolerance less than t, there is no standard value which complies with this requirement operations (d) to (h) may be repeated for a higher value of V_B. This provides only a gradual increase in the value of t, however. An increase in V_B increases the nominal value of R_S, but a convenient value of R_S may correspond to an arbitrary value of V_B. Usually it is simplest to reduce the resistor tolerance.

EXAMPLE 3.1 Glow-discharge Shunt Stabilizer

A CV286 is to be used to stabilize a load of 2–6 mA at nominal 95 V. Calculate suitable values of supply voltage (V_B) and series resistor (R_S) and the tolerance on R_S to meet supply variations of +6%, −10%. Reliable striking required at full load current.

Following the design procedure,

(a) Set out the following data:

$V_M = 95$ V, $\quad V_{IG} = 110$ V

$I_{K(min)} = 2$ mA $\quad I_{K(max)} = 10$ mA

$I_{L(min)} = 2$ mA $\quad I_{L(max)} = 6$ mA

Fractional tolerances on V_B, r_1 (positive tolerance) $= 0{\cdot}06$

r_2 (negative tolerance) $= 0{\cdot}10$

" " " V_M $\quad s_1 = s_2 = 0{\cdot}05$

(b) Evaluate

$q = V_{IG}/(1 - s_2)V_M = 110/0{\cdot}95 \times 95 = \mathbf{1{\cdot}22}$

$$n = \frac{[I_{K(max)} + I_{K(min)}] + [I_{L(max)} + I_{L(min)}]}{[I_{K(max)} - I_{K(min)}] - [I_{L(max)} - I_{L(min)}]} = \frac{10 + 2 + 6 + 2^*}{(10 - 2) - (6 - 2)} = 5$$

(c) Substitute values in Relation (3.3a):

$$V_B \geqslant \frac{V_{IG}}{(1 - r_2)} \cdot \frac{I_{K(max)} + I_{L(min)} - I_{L(max)}}{I_{K(max)} + I_{L(min)} - q \,.\, I_{L(max)}} \qquad (3.3a)$$

$$= \frac{110}{0{\cdot}90} \times \frac{10 + 2 - 6}{10 + 2 - 1{\cdot}22 \times 6} = \mathbf{157\ V}$$

Hence, to ensure full-load striking, even with no tolerance on R_S, V_B must exceed 157 V.

(d) Select a convenient value of V_B greater than the value of 157 V given by (c) above. Tentatively adopt a value of $V_B = 250$ V as convenient.

(e) Substitute in Relation (3.7) to determine the permissible resistor tolerance, t.

* Working in volts, milliamperes, and kilohms.

(put $r = \frac{1}{2}(r_1 + r_2) = 0{\cdot}08$)

$$t \leqslant \frac{\left(\frac{V_B}{V_M} - 1\right) - \left(\frac{r \,.\, V_B}{V_M} + s\right) . n}{\left(\frac{V_B}{V_M} - 1\right) . n - \left(\frac{r \,.\, V_B}{V_M} + s\right)} \qquad (3.7)$$

$$= \frac{\left(\frac{250}{95} - 1\right) - \left(\frac{0{\cdot}80 \times 250}{95} + 0{\cdot}05\right) \times 5}{\left(\frac{250}{95} - 1\right) \times 5 - \left(\frac{0{\cdot}08 \times 250}{95} + 0{\cdot}05\right)} = \mathbf{+0{\cdot}042}$$

(f) The resistor tolerance given by (e), although positive and therefore practicable, is too close to permit the use of a 5% resistor for R_S. Assuming this to be desired, the value of V_B is now tentatively raised to 300 V and operation (e) is repeated to obtain a higher value of t.

Thus, for $V_B = 300$ V, $t \leqslant \mathbf{+0{\cdot}061}$

(g) Use Equation (3.4c) to calculate the nominal value of R_S.

$$R_S = \frac{2\left\{\left(1 + \frac{r_1 - r_2}{2}\right) . V_B - V_M\right\}}{[I_{K(max)} + I_{K(min)} + I_{L(max)} + I_{L(min)}]} \qquad (3.4c)$$

$$= \frac{2\left\{\left(1 + \frac{0{\cdot}06 - 0{\cdot}10}{2}\right) \times 300 - 95\right\}}{(10 + 2 + 6 + 2)} = 20{\cdot}1\ \text{k}\Omega$$

(h) Select a standard resistance value of such a tolerance that the extreme values will lie within the limits 20·1 kΩ ± 6·1%. A satisfactory choice is 20 kΩ ± 5%.

Solution: $V_B = 300$ V, $R_S = 20$ kΩ ± 5%.

EXAMPLE 3.2 Corona Shunt Stabilizer

Design a circuit according to Fig. 3.3 to provide a stabilized output of 0–200 μA at 1 kV. Allow for supply variations of +10%, −15%. Base design on the G.E.C. corona stabilizer, Type S.C.1/1000.

(a) $V_M = 1{,}000 \quad I_{K(max)} = 0{\cdot}650$ mA $\quad r_1 = 0{\cdot}10$

$V_{IG} = 1{,}000 \quad I_{K(min)} = 0{\cdot}014$ mA $\quad r_2 = 0{\cdot}15$

$I_{L(max)} = 0{\cdot}200$ mA $\quad s_1 = 0{\cdot}025$

$I_{L(min)} = 0 \quad s_2 = 0{\cdot}025$

(b) $$q = \frac{1{,}000}{0{\cdot}975 \times 1{,}000} = \mathbf{1{\cdot}025}$$

$$n = \frac{(0{\cdot}650 + 0{\cdot}014) + (0{\cdot}200 + 0{\cdot}000)}{(0{\cdot}650 - 0{\cdot}014) - (0{\cdot}200 - 0{\cdot}000)} = \mathbf{1{\cdot}98}$$

(c) $$V_B \geqslant \frac{1000}{0{\cdot}85} \times \frac{0{\cdot}650 + 0 - 0{\cdot}200}{0{\cdot}650 + 0 - 1{\cdot}025 \times 0{\cdot}200} = \mathbf{1190\ V}$$

(*d*) Put $V_B = 1{,}500$ V

(*e*) Putting $r = r_2$,

$$t \leqslant \frac{\left(\frac{1500}{1000} - 1\right) - \left(0{\cdot}15 \times \frac{1500}{1000} + 0{\cdot}025\right) \times 1{\cdot}98}{\left(\frac{1500}{1000} - 1\right) \times 1{\cdot}98 - \left(0{\cdot}15 \times \frac{1500}{1000} + 0{\cdot}025\right)} = \mathbf{0{\cdot}0078}$$

(*g*)
$$R_S = \frac{2\left\{\left(1 + \frac{0{\cdot}10 - 0{\cdot}15}{2}\right) \times 1500 - 1000\right\}}{0{\cdot}650 + 0{\cdot}014 + 0{\cdot}200 + 0{\cdot}000} = \mathbf{1{\cdot}063\ k\Omega}$$

(*h*) R_S must lie within the limits of $1{\cdot}063$ MΩ, $\pm 7{\cdot}8\%$.
Hence put $R_S = 1{\cdot}047$ MΩ, $\pm 6\%$.
Solution: $V_B = 1{,}500$ V, $+10\%$, -15%.
$R_S = 1$ MΩ, $\pm 5\%$ in series with 47 kΩ, $\pm 20\%$.

Stabilization Factor

The maintaining voltage of a stabilizer tube is not quite independent of the tube current, even in the current range (which the manufacturer's recommended current range never exceeds) corresponding to normal glow. Over this range, the maintaining voltage increases gradually with increasing current. The slope of the corresponding characteristic is variously known as the incremental resistance (R_I), or impedance, of the tube. This parameter is important because it allows one to calculate the stabilization factor, *S*, of a circuit, i.e. the factor by which the supply variations are reduced.

In the circuit shown in Fig. 3.3 the stabilization factor, *S*, is given by:

$$S = 1 + \frac{R_S \, . \, R_L}{R_I(R_S + R_L)} \tag{3.8}$$

Hence if

$$R_L \gg R_S \gg R_I,$$
$$S \gtrsim R_S/R_I \tag{3.8a}$$

For stabilizers having maintaining voltages in the range 50–150 V, R_I is usually between 100 and 1,000 Ω. The series resistor, R_S, usually has a value of a few tens of kilohms, and *S* is thus commonly in the range of 20–500. Typically, supply variations amounting to several tens of volts may be reduced to several tenths of a volt.

Impedance

Variations in load applied across the tube may cause substantial tube current variations, and hence variations of a few volts in the tube maintaining voltage. Such voltage variations are, however, systematic, and the tube may be considered as a voltage source having an internal

impedance with resistive and inductive components. These have been studied by Benson [3, 4], who concludes [4] 'the complete equivalent circuit is much more complicated than the one usually quoted'. The inductive component arises because a change in tube current cannot instantaneously produce the changes in ionization and charge distribution necessary to keep the maintaining voltage approximately constant.

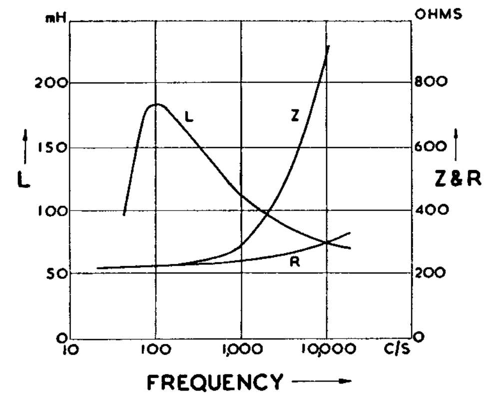


Fig. 3.6. Impedance/frequency characteristics of CV284.

Fig. 3.6 shows typical inductance, impedance, and resistance curves for a CV284. Although the inductance reaches a maximum at about 100 c/s, its effect is not significant below about 500 c/s. There is a relatively small change in the value of the resistive component with frequency.

As seen by a load connected in shunt with a voltage stabilizer, the resistive component, R_Z, of the equivalent source impedance is given by:

$$R_Z = \frac{R_I \, . \, R_S}{R_I + R_S} \, . \tag{3.9}$$

Or, when

$$R_S \gg R_I,$$
$$R_Z \gtrsim R_I \tag{3.9a}$$

Thus a change, δI_L, in the load current, I_L, produces a change, δV_M, in the maintaining voltage such that

$$\delta V_M = \delta I_L \, . \, R_Z \gtrsim \delta I_L \, . \, R_I \tag{3.10}$$

EXAMPLE 3.3 Regulation of Shunt Stabilizer

In the circuit described in Example 3.1 (p. 28) calculate the maximum voltage variations across the load (*a*) due to variations in supply voltage,

V_B, and (*b*) due to variations in load current, I_L. (Incremental resistance, R_I, of CV286 is 625 Ω.)

(*a*) From Equation (3.8) the smallest value of S occurs when R_L is a minimum, i.e. I_L is a maximum.

Now,
$$R_{L(min)} = \frac{V_M}{I_{L(max)}} = \frac{95}{6} = 15{\cdot}8 \text{ k}\Omega$$
from Equation (3.8),
$$S_{(min)} = 1 + \frac{20 \times 15{\cdot}8}{0{\cdot}625 \times 35{\cdot}8} = \mathbf{15{\cdot}1}$$

(This value is unusually low because the stabilizer is operating under extreme conditions.)

$$\text{Supply variation, } \delta V_B = (r_1 + r_2) \,.\, V_B = 0{\cdot}16 \times 300 = 48 \text{ V}$$
$$\text{Consequent variation across load} = \delta V_B/S = 48/15{\cdot}1 = \mathbf{3{\cdot}2\ V}$$

(*b*) Range of load current, $\delta I_L = 4$ mA

∴ from Equation (3.10) load variations produce a change, δV_M, in V_M, where:
$$\delta V_M \approx \delta I_L \,.\, R_I = 4 \times 0{\cdot}625 = \mathbf{2{\cdot}5\ V}$$

Fig. 3.6 shows that the impedance of a typical stabilizer rises abruptly above about 1 kc/s. This can lead to waveform distortion in pulse circuits, even though they may be operating at quite low frequencies. Often, however, the output impedance of the stabilizer circuit can be kept sufficiently low by connecting a capacitor in parallel with the tube. Usually 0·1 μF is sufficient to prevent a significant increase in impedance above that at 50 c/s. Relaxation oscillations (p. 38) or excessive peak currents on striking may occur if the shunt capacitance exceeds 0·5 μF.

Voltage Jumps

In early designs of voltage stabilizer the glow discharge could often be seen to jump abruptly from one part of the cathode area to another. Such jumps were accompanied by an abrupt change in maintaining voltage. Usually they occurred when the tube current changed, but sometimes they would apparently arise spontaneously. Designers of modern tubes try to minimize such voltage jumps or arrange for them to occur outside the working range of tube current.

A stabilizer tube is usually subject to fairly large current variations.

These produce variations in tube maintaining voltage of the order of a few volts (see Example 3.3). Voltage jumps of less than 1 V are thus not very important as a rule. If, however, an attempt is made to achieve improved stability by cascading stabilizer tubes as in Fig. 3.7, voltage jumps in V_2 may become significant compared with residual effects of supply variations.

Voltage Reference Tubes

Whereas a stabilizer tube provides a low-impedance source for variable loads, there are applications requiring the best possible stability for use in conjunction with a constant high-impedance load. To meet this need, a special type of tube has been developed in which the aims have been to minimize voltage jumps over a restricted current range, to improve long-term stability (at the price of an increased incremental resistance), and to minimize the temperature coefficient.

Cold cathode diodes of this type are known as voltage reference tubes. Typical examples have pure molybdenum electrodes from which a film of molybdenum is sputtered on to the inside of the envelope during manufacture. This film inhibits out-gassing of the glass and so ensures that the gas filling retains its initial high purity.

A voltage reference tube must be operated at substantially constant current if the relatively high incremental resistance is not to prove a limitation. It is thus standard practice to operate the reference tube

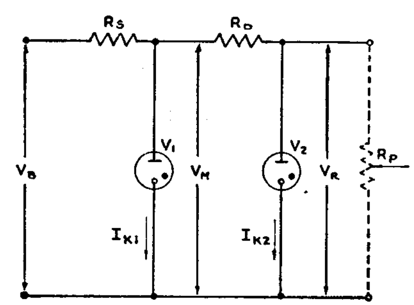


Fig. 3.7. Cascaded stabilizers.

from a supply controlled by a voltage stabilizer tube, as in Fig. 3.7. Variations in the supply, V_B, produce changes in V_M of the order of 1%. The dropping resistor, R_D, potentiometer, R_P, and reference tube, V_2, then reduce the voltage variations further still so that a short-term stability of better than 0·1% is obtained.

For design purposes that part of the circuit to the right of V_1 in Fig. 3.7 may be regarded as a constant load, R_L, where:

$$R_L = \frac{V_M}{V_M - V_R} \cdot R_D \qquad (3.11)$$

Design of a cascade arrangement of stabilizers thus entails determination of:

(i) appropriate values of V_M and R_D to suit V_R and R_P;
(ii) corresponding values of V_B and R_S.

In each case the 'worst × worst' condition should strictly be considered. This is too often overlooked, with the result that a value of V_M is chosen which is too low to ensure optimum working of V_2 without selection of tubes or adjustment of R_D.

Determination of V_M follows the pattern set for the determination of V_B in the case of the single stabilizer. There is this difference, however: it is usually the designer's object to keep V_M as low as possible, since the stabilizer must itself be supplied from a higher voltage. A low value of V_M implies a close tolerance, w, on the dropping resistor, R_D. The value of w may be determined by replacing V_B, V_M, I_L, r, s, and t in Relation (3.7) by V_M, V_R, I_P, s, v, and w respectively. Thus:

$$w \leqslant \frac{\left(\frac{V_M}{V_R} - 1\right) - \left(\frac{s \cdot V_M}{V_R} + v\right) \cdot n}{\left(\frac{V_M}{V_R} - 1\right) \cdot n - \left(\frac{s \cdot V_M}{V_R} + v\right)} \qquad (3.12)$$

where $m = \dfrac{[I_{K2(max)} + I_{K2(min)}] + [I_{P(max)} + I_{P(min)}]}{[I_{K2(max)} - I_{K2(min)}] - [I_{P(max)} - I_{P(min)}]}$,

s = fractional tolerance on V_M,
v = fractional tolerance on V_R,
w = fractional tolerance on R_D.

Design Procedure for Cascaded Stabilizers

(*a*) Set out the following data:

Nominal reference tube maintaining voltage, V_R
Maximum reference tube ignition voltage, V_{IG2}
Minimum and maximum reference tube currents, $I_{K2(min)}$ and $I_{K2(max)}$, (relating to range within which reference tube is to operate)
Nominal load resistance, R_P, in shunt with reference tube
Fractional tolerance, $\pm p$, on the load resistance, R_P

Fractional tolerances, $+r_1$ and $-r_2$, on supply voltage, V_B
Fractional tolerance, $\pm s$, on the stabilizer tube voltage, V_M*
Fractional tolerance, $\pm v$, on the reference tube voltage, V_R.

(*b*) Evaluate

$$q_2 = \frac{V_{IG2}}{(1 - v)V_R}$$

$$I_{P(min)} = \frac{(1 - v) \cdot V_R}{(1 + p) \cdot R_P}$$

$$I_{P(max)} = \frac{(1 + v) \cdot V_R}{(1 - p) \cdot R_P}$$

$$m = \frac{[I_{K2(max)} + I_{K2(min)}] + [I_{P(max)} + I_{P(min)}]}{[I_{K2(max)} - I_{K2(min)}] - [I_{P(max)} - I_{P(min)}]}$$

(*c*) Substitute values in Relation (3.3b) to determine the minimum value of stabilizer maintaining voltage, V_M, required to ensure striking of V_2.

$$V_M \geqslant \frac{V_{IG2}}{(1 - s)} \cdot \frac{I_{K2(max)} + I_{P(min)} - I_{P(max)}}{I_{K2(max)} + I_{P(min)} - q_2 \cdot I_{P(max)}} \qquad (3.3b)$$

(*d*) Tentatively select a stabilizer having a nominal maintaining voltage in excess of that given by (*c*). Set out relevant data.

(*e*) Evaluate the limiting tolerance, w, on the dropping resistor, R_D, given by Relation (3.12).

$$w \leqslant \frac{\left(\frac{V_M}{V_R} - 1\right) - \left(\frac{s \cdot V_M}{V_R} + v\right) \cdot m}{\left(\frac{V_M}{V_R} - 1\right) \cdot m - \left(\frac{s \cdot V_M}{V_R} + v\right)} \qquad (3.12)$$

(*f*) (i) If the value of w so obtained is negative or too small, repeat operations (*d*) and (*e*) for another tube type (or series combination of tubes) giving a higher value of V_M.

(ii) When a satisfactory value of w is given by Relation (3.12), evaluate R_D from Equation (3.4d)

$$R_D = \frac{2(V_M - V_R)}{[I_{K2(max)} + I_{K2(min)} + I_{P(max)} + I_{P(min)}]} \qquad (3.4d)$$

(iii) Adopt convenient values for R_D and w such that the extreme values of R_D lie within the limits imposed by the calculated tolerance. Hereafter work with the convenient (closer) tolerance, w.

* This is not definitely known until the value of V_M has been settled.

(*g*) Use Equation (3.4e) to calculate the minimum load current, $I_{L(min)}$, in shunt with V_1.

$$I_{L(min)} = \frac{(1-s)\,.\,V_M - (1+v)\,.\,V_R}{(1+w)\,.\,R_D} \tag{3.4e}$$

(*h*) Similarly, use Equation (3.4f) to calculate the maximum load current, $I_{L(max)}$, in shunt with V_1.

$$I_{L(max)} = \frac{(1+s)\,.\,V_M - (1-v)\,.\,V_R}{(1-w)\,.\,R_D} \tag{3.4f}$$

(*j*) Substitute in Relation (3.3a) to determine the minimum value of V_B ensuring striking of V_1.

$$V_B \geqslant \frac{V_{IG1}}{(1-r_2)} \cdot \frac{I_{K1(max)} + I_{L(min)} - I_{L(max)}}{I_{K1(max)} + I_{L(min)} - q_1\,.\,I_{L(max)}} \tag{3.3a}$$

(*k*) Evaluate

$$n = \frac{[I_{K1(max)} + I_{K1(min)}] + [I_{L(max)} + I_{L(min)}]}{[I_{K1(max)} - I_{K1(min)}] - [I_{L(max)} - I_{L(min)}]}$$

(*l*) Tentatively choose a value of V_B in accordance with the value obtained in (*j*) above.

(*m*) For the chosen value of V_B, evaluate the maximum corresponding tolerance, t, on R_S as given by Relation (3.7).

$$t \leqslant \frac{\left(\dfrac{V_B}{V_M} - 1\right) - \left(\dfrac{r\,.\,V_B}{V_M} + s\right).\,n}{\left(\dfrac{V_B}{V_M} - 1\right).\,n - \left(\dfrac{r\,.\,V_B}{V_M} + s\right)} \tag{3.7}$$

(If $r_1 \neq r_2$, a mean value may be taken for r if the two values are not greatly different. Alternatively, a factor of safety may be added by adopting the larger value.)

(*n*) (i) If the value obtained for t is negative, repeat operations (*l*) and (*m*) for a higher value of V_B.

(ii) When an acceptable (positive) value of t is obtained, calculate from Equation (3.4g) the nominal value of R_S.

$$R_S = \frac{2\left\{\left(1 + \dfrac{r_1 - r_2}{2}\right).\,V_B - V_M\right\}}{[I_{K1(max)} + I_{K1(min)} + I_{L(max)} + I_{L(min)}]} \tag{3.4g}$$

EXAMPLE 3.4 Cascaded Shunt Stabilizer

Design a cascade arrangement of stabilizer and reference tube (CV449) to operate from an unstabilized supply subject to voltage variations of

+6%, −10%. A load of 47 kΩ, ±1% is to be connected across the CV449.

Following the design procedure:

(*a*) Set out $V_R = 85$ V, $V_{IG2} = 115$ V.

Permissible operating range of CV449 is 1 – 10 mA, but for good performance the current should be close to 6 mA.

Put: $I_{K2(max)} = 7$ mA, $p = 0{\cdot}01$, $s = 0{\cdot}027$*
$I_{K2(min)} = 5$ mA, $r_1 = 0{\cdot}06$, $v = 0{\cdot}024$
$R_P = 47$ k, $r_2 = 0{\cdot}10$

(*b*) Evaluate:

$$q_2 = \frac{115}{0{\cdot}976 \times 85} = \mathbf{1{\cdot}39}$$

$$I_{P(min)} = \frac{(1 - 0{\cdot}024) \times 85}{(1 + 0{\cdot}010) \times 47} = 1{\cdot}75 \text{ mA}$$

$$I_{P(max)} = \frac{(1 + 0{\cdot}024) \times 85}{(1 - 0{\cdot}010) \times 47} = 1{\cdot}87 \text{ mA}$$

$$m = \frac{(7+5) + (1{\cdot}87 + 1{\cdot}75)}{(7-5) - (1{\cdot}87 - 1{\cdot}75)} = 8{\cdot}32$$

(*c*) Substituting values in Relation (3.3b),

$$V_M \geqslant \frac{115}{0{\cdot}973} \times \frac{7{\cdot}00 + 1{\cdot}75 - 1{\cdot}87}{7{\cdot}00 + 1{\cdot}75 - 1{\cdot}39 \times 1{\cdot}87} = \mathbf{132\ V}$$

(*d*) Tentatively select CV2225 as the stabilizer and set out:

$V_M = 150$ V, $V_{IG1} = 180$ V, $q_1 = 180/0{\cdot}90 \times 150 = \mathbf{1{\cdot}34}$
$I_{K1(min)} = 5$ mA, $I_{K1(max)} = 15$ mA, $s = 0{\cdot}027$

(*e*) From Relation (3.12), evaluate the limiting tolerance, w, on the dropping resistor, R_D.

$$w \leqslant \frac{\left(\dfrac{150}{85} - 1\right) - \left(0{\cdot}27 \times \dfrac{150}{85} + 0{\cdot}024\right) \times 8{\cdot}32}{\left(\dfrac{150}{85} - 1\right) \times 8{\cdot}32 - \left(0{\cdot}027 \times \dfrac{150}{85} + 0{\cdot}024\right)} = \mathbf{0{\cdot}027}$$

(Note that to restrict I_{K2} to ±16% of 6 mA it is necessary to impose on R_D a tolerance of only ±2·7%.)

(*f*) (ii) $R_D = \dfrac{(150 - 85)}{(6 + 1{\cdot}81)} = \mathbf{8{\cdot}32\ k\Omega}$

(iii) By calculation, $R_D = 8{\cdot}32$ kΩ ± 2·7%. To use a standard value, put $R_D = 8{\cdot}2$ kΩ ± 1% ($w = 0{\cdot}010$) and hereafter work to these values.

* This value of s assumes tentatively that a CV2225 will be used as the stabilizer.

(*g*) $$I_{L(\min)} = \frac{(0{\cdot}973 \times 150 - 1{\cdot}024 \times 85)}{1{\cdot}010 \times 8{\cdot}20} = \mathbf{7{\cdot}13\ mA}$$

(*h*) $$I_{L(\max)} = \frac{(1{\cdot}027 \times 150 - 0{\cdot}976 \times 85)}{0{\cdot}99 \times 8{\cdot}20} = \mathbf{8{\cdot}74\ mA}$$

(*j*) $$V_B \geqslant \frac{180}{0{\cdot}90} \times \frac{15{\cdot}0 + 7{\cdot}13 - 8{\cdot}74}{15{\cdot}0 + 7{\cdot}13 - 1{\cdot}34 \times 8{\cdot}74} = \mathbf{256\ V}$$

(*k*) $$n = \frac{(15 + 5) + (8{\cdot}74 + 7{\cdot}13)}{(15 - 5) - (8{\cdot}74 - 7{\cdot}13)} = \mathbf{4{\cdot}28}$$

(*l*) Tentatively put $V_B = 300$ V

(*m*) $$t \leqslant \frac{\left(\frac{300}{150} - 1\right) - \left(0{\cdot}08 \times \frac{300}{150} + 0{\cdot}027\right) \times 4{\cdot}28}{\left(\frac{300}{150} - 1\right) \times 4{\cdot}28 - \left(0{\cdot}08 \times \frac{300}{150} + 0{\cdot}027\right)} = \mathbf{0{\cdot}049}$$

(*n*) (ii) $$R_S = \frac{2\left\{\left(1 + \frac{0{\cdot}06 - 0{\cdot}10}{2}\right) \times 300 - 150\right\}}{(15{\cdot}0 + 5{\cdot}0 + 8{\cdot}74 + 7{\cdot}13)} = \mathbf{8{\cdot}03\ k\Omega}\ (\pm 4{\cdot}9\%)$$

To use a standard value, put $R_S = 8{\cdot}2$ kΩ, ±2%.

Solution: V_1 = CV2225, $V_B = 300$ V, $R_S = 8{\cdot}2$ kΩ ± 2%,
$R_D = 8{\cdot}2$ kΩ ± 1%.

Relaxation Oscillations

Under normal circumstances the discharge through a diode can be extinguished only by reducing the supply voltage below the maintaining voltage. This may be done by interrupting the supply or by shunting the tube sufficiently heavily. A discharge may be made self-quenching,

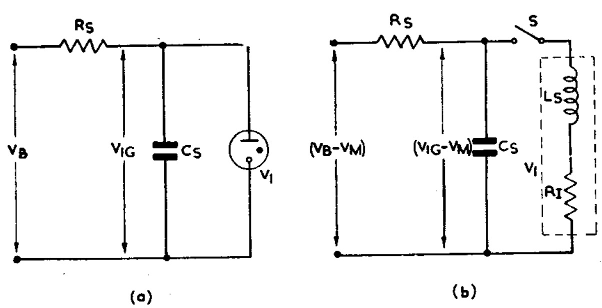


Fig. 3.8. Relaxation oscillator. (*a*) Practical circuit, and (*b*) simplified equivalent circuit.

however, by connecting a substantial capacitance, C_S, in shunt with the tube, as in Fig. 3.8 (*a*). When V_B exceeds the tube ignition voltage, V_1 strikes and then extinguishes almost immediately. This is to be expected if it is remembered that the tube impedance contains a large inductive component at low frequencies. The striking of V_1 in Fig. 3.8 (*a*) is thus analogous to closure of the switch S in Fig. 3.8 (*b*). A resonant circuit is formed by L_I and C_S. If the damping by R_I and R_S is less than critical the step-function ($V_{IG} - V_M$) produces a voltage oscillation about the normal maintaining voltage, V_M. The voltage across the tube thus swings below V_M, the current falls to zero, and the discharge ceases.

This analogy cannot profitably be followed very far because the equivalent inductance of the tube is a function of frequency. It can, however, be seen that if R_S is made sufficiently low the resonant circuit will suffer more than critical damping. In such cases the tube voltage does not swing below V_M, and thus the discharge does not quench. For the reason stated, a theoretical analysis is impracticable and for a given value of C_S the critical value of R_S is best determined by experiment. As a general rule, it may be taken that self-quenching is likely if R_S exceeds 1 MΩ and unlikely if R_S is less than 100 kΩ. Larger values of C_S produce self-quenching more readily than small ones.

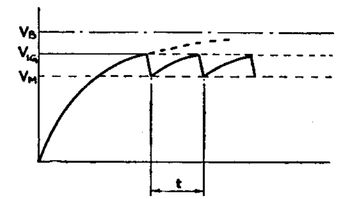


Fig. 3.9. Waveform of relaxation oscillations.

When the tube current is self-quenching C_S is left at a potential slightly below V_M. Thereafter current flows through R_S to recharge C_S exponentially towards V_B. When V_{IG} is again reached, however, V_1 again strikes and again quenches. Relaxation oscillations thus occur characterized by the saw-toothed voltage waveform depicted in Fig. 3.9.

If the time taken to discharge C_S from V_{IG} to V_M is neglected as being relatively small the periodic time, t, of the relaxation oscillation may be calculated as the difference between the times t_1 and t_2 required to charge C_S to V_M and V_{IG} respectively.
Thus:

$$t = t_1 - t_2$$

$$= \left[-CR \log_e \frac{(V_B - V_{IG})}{V_B}\right] - \left[-CR \log_e \frac{(V_B - V_M)}{V_B}\right]$$

$$= CR \log_e \frac{(V_B - V_M)}{(V_B - V_{IG})} \qquad (3.13)$$

In practice, the frequency of oscillation will be lower than the calculated value of $1/t$ because the discharge time may be only one order less than the periodic time. Also the capacitor is discharged not to V_M, as Equation (3.13) assumes, but to a slightly lower voltage.

Stabilizers and voltage reference tubes having pure metal cathodes can have highly stable maintaining voltages. Their ignition voltages are not very stable, however, and breakdown is subject to statistical delays. Consequently, at sub-audio frequencies relaxation oscillators of this type do not provide good frequency stability. Even tubes containing tritium are not exempt from this criticism. Above about 50 *c/s* deionization between discharges is usually incomplete, and better frequency stability is thus obtained.

Diodes having oxide-coated cathodes do not suffer from statistical delays, and so can perform fairly well at low frequencies. Ambient illumination is essential, however, if the breakdown voltage is to be reasonably low and stable. To some extent the level of illumination influences the breakdown potential, and hence the frequency of the oscillator.

The maximum frequency at which a diode oscillator can operate is determined by the deionization time of the tube. This varies from one tube to another, but it is seldom possible to exceed an upper frequency limit of a few kilocycles per second.

Neon diode relaxation oscillators have been used in electronic organs, as they are cheap and provide output rich in harmonics. Douglas [5] has discussed the difficulty of attaining good frequency stability with inexpensive diodes.

A multi-diode ring described by German [6] is a form of relaxation oscillator. This circuit appears to be no more than an amusing toy, although it has much in common with the ring counter described below.

Deionization Time

A discharge may be extinguished by holding the anode for a suitable time below the normal maintaining voltage, V_M. The deionization time has been defined by Hough and Ridler [7] as 'the time which must elapse following a d.c. discharge of maximum rated current, extinguished by a rectangular pulse applied to the anode, before 90% of the maximum working voltage may be re-applied across the gap without re-igniting the discharge'. This definition was framed in relation to trigger tubes, but it is equally applicable to diodes if one takes 'maximum working voltage' to indicate 'ignition voltage'.

An alternative definition of deionization time is implicit in the method

of determination described by Hendrix [13]. This is based on the observation that an incompletely deionized tube will restrike at a voltage lower than its normal ignition voltage. Hendrix determines the lowest frequency of square wave supply with which the restriking voltage is measurably lower than the ignition voltage of the fully deionized tube. The 'off' period of the supply waveform is then concluded to be just less than the deionization time of the tube.

The deionization time depends very largely on the voltage at which the anode is held while deionization takes place. If the anode is held only a few volts below the maintaining voltage the field is only just insufficient to maintain the discharge and deionization is slow, being measured in milliseconds. Deionization is about equally slow if the anode voltage is entirely removed, since under these circumstances there is no field to effect removal of ions from the gap. At about half the maintaining voltage, however, the field is strong enough to remove ions quickly without producing significant further ionization. Under these circumstances deionization times of a few microseconds can be obtained with suitably designed tubes.

Much then depends on the nature of the electrodes and the gas filling. Since diodes are seldom used for purposes for which short deionization times are required, relevant figures are not generally available. Some guidance may, however, be obtained from the notes on deionization time in trigger tubes (p. 68). These indicate that tubes having oxide-coated cathodes will show deionization times of the order of 1 or 2 msec. This is supported by measurements made by Hendrix [13] on the NE2 diode.

A tube carrying a small current is less heavily ionized than one carrying a large current. Accordingly, the tube operating at small current will deionize more quickly.

Circuit Applications

A number of circuits are based on the connexion in parallel of several similar diodes, supplied through a common series resistor from a supply voltage greater than the ignition voltage. In such an arrangement one of the tubes must strike. As soon as this happens, however, the voltage across the other tubes falls to the maintaining voltage of the conducting tube. Thus no other tubes can strike. If the supply voltage is briefly reduced below the maintaining voltage the conducting tube begins to deionize. As soon as the supply voltage returns to the striking voltage, however, the same tube will normally restrike.

In the circuits described below special features are obtained either

by (*a*) ensuring that the same tube does not restrike, or by (*b*) ensuring that a particular (alternative) tube strikes in preference to any other. With more than two tubes in parallel the first alternative becomes ambiguous.

Scale-of-two

When only two similar tubes are connected in parallel, inhibiting the striking of one tube is almost equivalent to promoting striking of the other. The circuit of Fig. 3.10 is thus the basis of a flip-flop, binary counter, or scale-of-two.

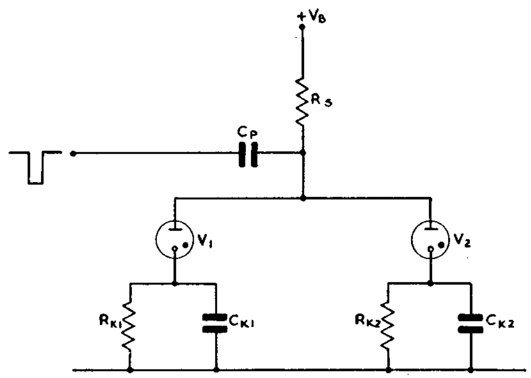


Fig. 3.10. Neon diode scale-of-two.

The supply voltage, V_B, is greater than the striking voltage, V_{IG}, of V_1 or V_2. Thus one or other of the tubes will normally be conducting. Assume this is V_1. Then the current I_{K1} through V_1 will be given by:

$$I_{K1} = \frac{(V_B - V_M)}{(R_{K1} + R_S)} \qquad (3.14)$$

and C_{K1} will be charged to a voltage $I_{K1}R_{K1}$. The voltage appearing across V_2 therefore rises to $(V_M + I_{K1}R_{K1})$, and provided this does not exceed V_{IG}, V_2 will not strike.

If, now, a negative-going pulse is applied through the capacitor C_P to the anodes of V_1 and V_2 the voltage across V_1 will momentarily fall below V_M. Deionization of V_1 then begins, and simultaneously C_{K1} begins to discharge through R_{K1}. If the pulse terminates after V_1 has deionized but before C_{K1} has discharged significantly, both anodes will return to their original potential $(V_M + I_{K1}R_{K1})$. But whereas the cathode of V_2 will be at ground potential, the cathode of V_1 will still be

positive by a substantial proportion of the voltage $I_{K1}R_{K1}$. Thus a greater voltage appears across V_2 than across V_1. The anode potential of both tubes now rises at a rate determined by the time-constant C_PR_S. When the value V_{IG} is reached V_2 strikes and the anode potential quickly returns to its normal steady-state value. C_{K2} charges to the voltage $I_{K2}R_{K2}$ $(= I_{K1}R_{K1})$ and C_{K1} discharges through R_{K1}. By similar processes, the condition of the circuit reverts to its original state on receipt of the next input pulse via C_P.

An output may be taken from the cathode of either V_1 or V_2 by

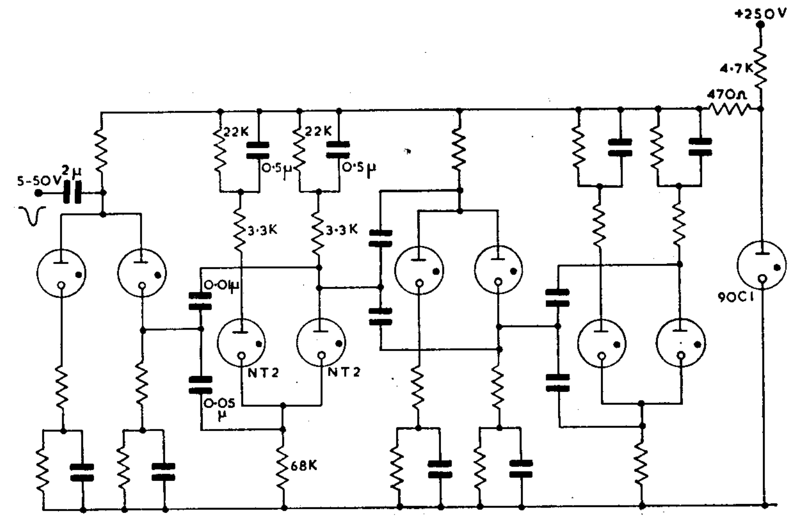


Fig. 3.11. Binary scaler using neon diodes.

either a.c. or d.c. coupling. A resistor in series with C_{K2} provides a differentiated output. If the resistor is shunted by a diode only positive-going pulses need appear, corresponding to the striking of V_2 at the end of alternate input pulses. Using an a.c. coupling of appropriate time-constant, these output pulses may be taken to a further scale-of-two of similar design. Since the output from the first stage is positive-going, however, it is necessary to 'invert' the second stage. Any number of such stages may be cascaded, provided they are alternately inverted.

Fig. 3.11 shows a complete binary scaler of this type.

The maximum counting rate of a diode binary scaler is not very high – perhaps 100–200 c/s. To minimize delays in striking, tubes must be used having oxide-coated cathodes, and these must be illuminated. Unfortunately such tubes have relatively long deionization times. To obtain

the maximum counting rate from a given tube, the input pulse should ideally be rectangular and of an amplitude equal to about half the tube maintaining voltage. (Neither of these requirements is likely to be met in stages after the first, but as they operate more slowly this is not important.)

It is more usual to employ an exponential input pulse. This must be of a larger peak amplitude, and for maximum counting speed the time-constant should be one-third or one-quarter of the deionization time as defined above [7]. The maximum counting speed will nevertheless be inferior to that attainable with rectangular pulses.

The cathode time-constant $R_{K1}C_{K1}$ ($= R_{K2}C_{K2}$) should be about an order greater than that of the input pulse, i.e. about three times the nominal deionization time. If the counter is intended only for operation at low frequencies greater reliability can be obtained by increasing $R_{K1}C_{K1}$ and $R_{K2}C_{K2}$.

As the impedance of the conducting tube is relatively low, the input impedance is approximately equivalent to R_S, R_K, and C_K in parallel. Driving pulses of suitably low impedance may be obtained from a bounce-free switch or the striking of another cold cathode tube, as described later.

Vuylsteke [8] has described an alternative type of binary counter. This will operate with either positive- or negative-going input pulses. Four neon diodes are required per binary stage.

Diode Ring Counter

Instead of inhibiting restriking of the extinguished tube, striking of a particular alternative tube may be encouraged, and in this way a ring counter may be produced which will operate predictably with far more than two tubes in the ring. It proves most convenient to 'invert' alternate stages, and in consequence the ring must generally contain an even number of tubes. (This restriction does not apply to ring counters using trigger tubes and described in Chapter Four.)

Four successive stages of a diode ring counter due to Manley and Buckley [9] are shown in Fig. 3.12. Further stages may be added by repeating the pattern set by the first two. The ring is completed by coupling the last stage to the first.

Suppose V_0 is conducting. The voltage across the other tubes will be $(V_M + I_K R_K)$. If this is less than V_{IG} no other tubes will strike. Consider the effect of a negative-going input pulse applied to C_P. If the pulse amplitude exceeds $I_K R_K$ the voltage across V_0 falls below V_M and V_0 begins to deionize. No current now flows through V_0 and its

cathode is at ground potential. The charge on C_0 thus holds the cathode of V_1 at a negative potential of $I_K R_K$ and, since neither V_1 nor the diode D_1 conducts, C_0 cannot discharge. At the end of the input pulse the rail A returns to its original voltage, $(V_M + I_K R_K)$. This voltage appears across all tubes except V_1. Across this tube there is a greater

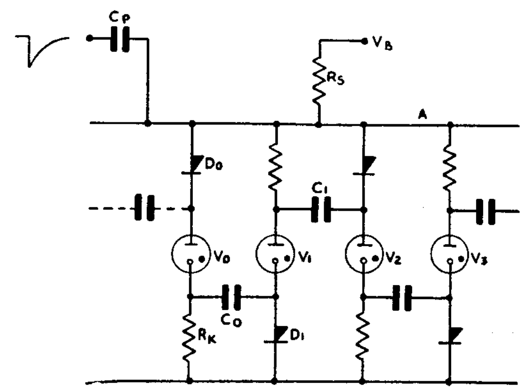


Fig. 3.12. Four stages of diode chain or ring counter.

potential difference due to the negative bias applied to its cathode by C_0. In fact, a voltage $(V_M + 2I_K R_K)$ is available to strike V_1 and if this is insufficient the potential of the rail A will rise further until V_1 strikes. As soon as V_1 strikes, D_1 conducts, and C_0 discharges through R_K.

When another input pulse is applied through C_P a similar mechanism ensures that V_1 extinguishes and C_1 promotes striking of V_2 at the end of the pulse.

From the foregoing, it will be seen that

$$[V_M + 2I_K R_K] > V_{IG} > [V_M + I_K R_K]$$

whence

$$R_K \approx \frac{2}{3} \cdot \frac{(V_{IG} - V_M)}{I_K} \tag{3.15}$$

In the absence of an input pulse, the tube current I_K is given by

$$I_K(R_K + R_S) = (V_B - V_M) \tag{3.16}$$

Hence, having chosen a tube type and a working current I_K, one can calculate R_K using Equation (3.15). A value of V_B is then chosen exceeding V_{IG} and R_S is determined from Equation (3.16).

To ensure that V_{IG} for one tube is never less than V_M for another tube, the circuit should be designed either for close-tolerance tubes or

for 'difference diodes', i.e. diodes with a relatively large difference between V_{IG} and V_M.

Provision may be made for reversing the direction of count by switching all inter-stage coupling capacitors so that transfer is effected

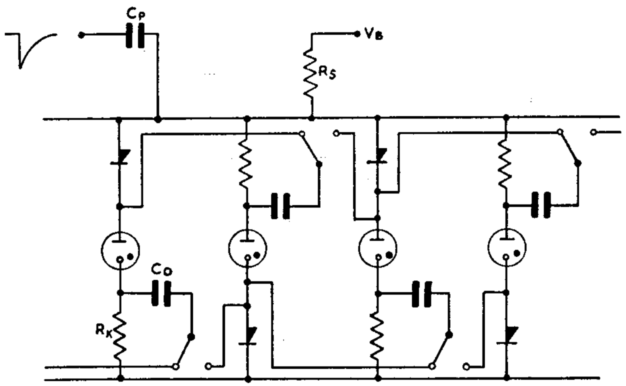


Fig. 3.13. Diode ring counter with ganged switches set to 'reverse' position.

from right to left instead of from left to right. The capacitor must remain connected to the resistor, rather than the semiconductor diode, if the state of count is not to be disturbed when the direction-reversing switch is operated. A suitable arrangement is shown in Fig. 3.13.

The value of the coupling time-constant, C_0R_K, must be such that

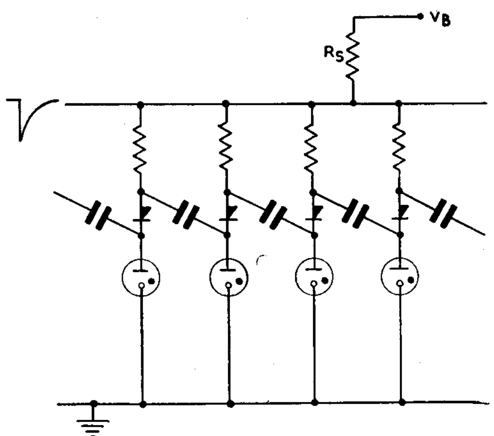


Fig. 3.14. Diode chain or ring counter in which all stages are identical.

C_0 is substantially discharged before the ring counter has gone through a complete cycle of n counts. The input time-constant, C_PR_S, should be not less than one-third of the tube deionization time – nor much more if rapid counting is required. Usually it is convenient to put $C_P \approx C_0$ and $R_S \approx R_K$.

Flood and Warman [10] have described a diode ring counter in which, by rearranging the same components as in Fig. 3.14, all stages become identical. The design considerations discussed above are still applicable.

Trigger Pulse Generation

Exponential trigger pulses may be generated by switch contacts using the circuit of Fig. 3.15 (*a*). While the contacts are open, point A charges to a fraction $R_a/(R_a + R_b)$ of the supply voltage. When S is closed A falls abruptly to ground potential and a negative-going pulse is applied through C_P to the input rail of the counter. The amplitude, V_P, of this pulse is given by

$$V_P = \frac{R_a}{R_a + R_b} \cdot V_B \qquad (3.17)$$

Figs. 3.15 (*b*) and (*c*) show methods of using a cold cathode diode to generate trigger pulses. These are useful for inter-decade coupling or

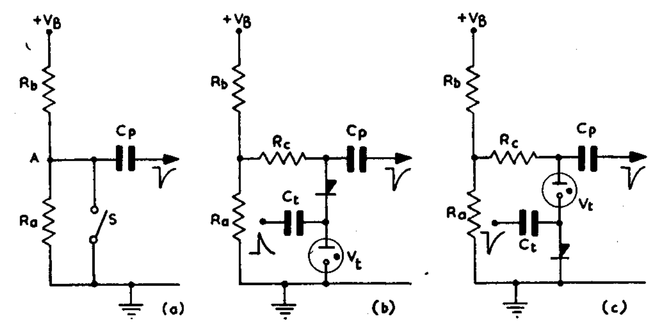


Fig. 3.15. Generation of exponential trigger pulses by (*a*) switch contacts, (*b*) positive pulse to diode, and (*c*) negative pulse to diode.

where input pulses are obtained from a high-impedance source such as a photocell. The ratio R_a/R_b is chosen so that the voltage applied to the tube is a little less than V_{IG}. A positive input pulse applied to C_t in Fig. 3.15 (*b*) – or a negative pulse in Fig. 3.15 (*c*) – momentarily cuts off the semiconductor diode and applies an increased voltage to the tube, thereby striking it. The voltage across the tube now falls abruptly

to the maintaining voltage, V_M, the semiconductor diode again conducts and a negative-going pulse is delivered through C_P. Its amplitude, V_P, is given by

$$V_P = \frac{R_a}{R_a + R_b} \cdot V_B - V_M \tag{3.18}$$

In the case of the binary and ring counters described already it is necessary to make $V_P > I_K R_K$. Hence, combining Equations (3.15) and (3.18),

$$\frac{R_a}{R_a + R_b} \cdot V_B > \frac{2}{3} V_{IG} \tag{3.19}$$

The standing voltage across the cold cathode diode must not exceed V_{IG}. Hence a reasonable compromise is usually found in

$$\frac{R_a}{R_a + R_b} V_B \approx 0{\cdot}9 V_{IG} \tag{3.20}$$

i.e.

$$\frac{R_b}{R_a} \approx 1{\cdot}1 \left(\frac{V_B}{V_{IG}}\right) - 1 \tag{3.20a}$$

The delivery of an output pulse from the circuit of Fig. 3.15 (*b*) or (*c*) extinguishes the conducting tube in the circuit to which it is connected.

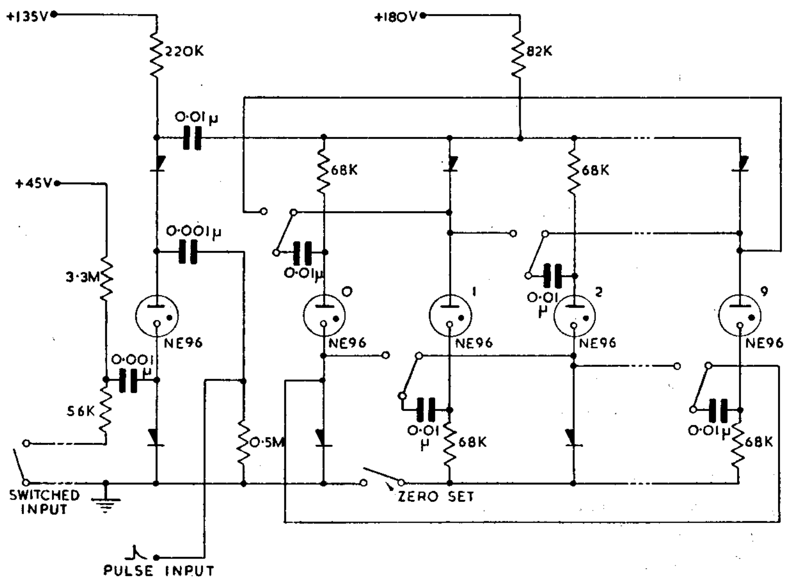


Fig. 3.16. Reversible diode ring decade counter. Tube at left generates stepping pulses in response to either switched or pulse inputs.

As the pulse decays, however, another tube strikes in the counting circuit. The common input rail of the counter falls by nearly ($V_{IG} - V_M$) and a negative-going pulse of this amplitude is transmitted by C_P back to the triggering tube, V_t. Thus the discharge in V_t is automatically extinguished.

The input pulse applied through C_t must terminate before the end of the quenching pulse returned through C_P. If this is not so, V_t will not be extinguished. It is usually sufficient to arrange that $C_t < C_P$. The time constant $C_P R_C$ is best made roughly equal to the deionization time of the counting tubes. Thus in a ring counter, $R_C \approx 3R_S$.

The decade counter shown in Fig. 3.16 is based on a circuit described by Manley and Buckley [9]. This circuit combines various techniques described above. In particular, it will be noted that the triggering arrangements shown in Fig. 3.15 (*b*) and (*c*) are combined so that triggering may be effected either by an input pulse or by switched input.

Logical and Memory Circuits

Neon diodes have been used as logical and storage elements for data-handling systems. Their value is restricted by their ionization and de-ionization times, but for low- and medium-speed applications they are attractive, since they are inexpensive and provide a visible indication of operation.

Neon Diode Shift Register

A simple pattern shifting register (Fig. 3.17) due to Warman uses two pulse inputs to circulate the charge 'written in' by closing switch S_1. Suppose S_1 is momentarily closed to charge C_1 to the potential V_B. Provided $V_B < V_{IG}$ neither V_1 nor V_8 will ionize and C_1 remains charged to V_B. When a positive pulse, V_{P2}, is applied to rail P_2, however, V_1 conducts to transfer charge from C_1 to C_2. A positive pulse, V_{P1}, applied to P_1 now causes V_2 to conduct so that the charge is transferred from C_2 to C_3. S_1 may now be closed to recharge C_1 if it is required to 'write in' another '1'. As pulses are alternately applied to P_2 and P_1, the stored charges progress first from C_{odd} to C_{even} and then from C_{even} to the next C_{odd}.

Flood and Warman [10] discuss the circuit operation in some detail and show that, to produce complete transfer of charge from one capacitor to the next,

$$V_P = V_B + V_M \tag{3.21}$$

In practice, V_M differs significantly from tube to tube, but this leads to the production of residual charges providing a compensating action.

As a result, the charge circulating does not depend on tube tolerances. Also, provided the stored charges are circulated, they do not diminish due to leakage effects. By comparison with a shift register using triodes

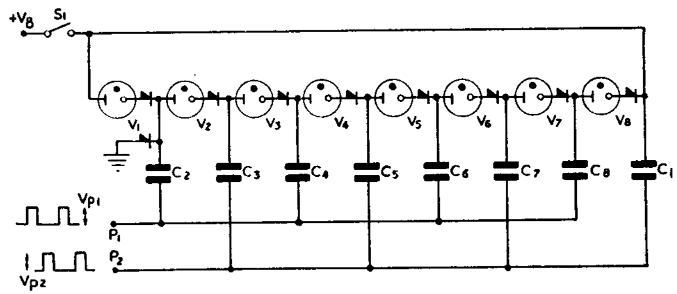


Fig. 3.17. Pattern shifting register.

(p. 143), the diode shift register provides rapid operation (at least up to 1,000 steps/sec) and low power consumption. On the other hand, charge leakage effects may become serious if the register is not stepped several times a minute at least. Also no visible indication of state is provided with the store static. The diode circuit uses two neon diodes and two semiconductor diodes per storage position, whereas the triode circuit uses only one trigger tube and three or four resistors in addition to the two capacitors per storage position.

Neon Diode Logical Matrix

A logical matrix using neon diodes has been described by Raphael and Robinson [11]. Tubes are connected between two sets of wires crossing at right angles, as shown in Fig. 3.18. Each tube is thus connected to one 'horizontal' and one 'vertical' wire, and the particular combination of wires is unique to that tube. At any instant, only one horizontal and one vertical wire is energized. If V_M is applied to a horizontal wire and $-V_M$ to a vertical wire a neon connected between these two wires will strike. A multiplier photocell receives light from any neon which strikes and so indicates a '1'. If there is no neon connected between the wires energized the photocell remains unilluminated and the absence of an output indicates a '0'. Complications arise because the neons and photomultiplier must be in an enclosure excluding stray light. In consequence, striking of a neon may be delayed due to the absence of primary electrons. This difficulty is overcome by using a 'flasher' neon tube to stimulate photo-emission not more than 30 msec before any matrix neon

is required to strike. The photomultiplier, of course, receives light from the flasher neon, but the unwanted output pulse thereby produced is removed by a gated amplifier.

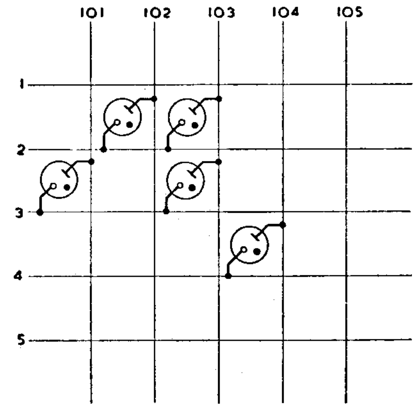


Fig. 3.18. Diode logical matrix.

'Dynamic Bit' Memory Circuit

Memory circuits due to Hold and Friedman [12] provide for the rapid 'writing' of new information into a memory. In Fig. 3.19 (*a*) a 5-μsec pulse is applied to the point A at intervals of 100 μsec. The pulse

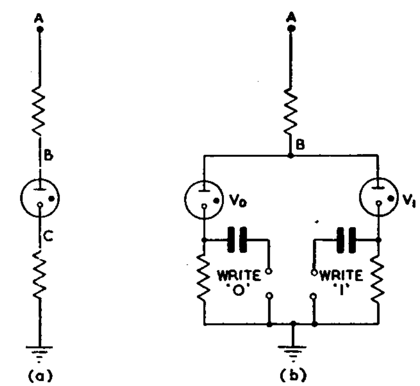


Fig. 3.19. 'Dynamic bit' memory circuits, (*a*) requiring 100 μsec to erase a '1', and (*b*) requiring only 5 μsec to write either '1' or '0'.

has an amplitude midway between V_M and V_{IG} so that if the tube is not already ionized this pulse is not capable of producing ionization. In this condition the tube is in the '0' state. A '1' is written into the tube by

applying simultaneously a positive pulse at B and a negative pulse at C. (For this purpose, each tube in a memory is connected across one intersection of a crossed-wire matrix similar to that in Fig. 3.18.) Such a combination of positive and negative pulses ionizes the tube, and ionization is therafter maintained by the pulses at A so long as these are repeated at 100 μsec intervals. On a tube storing a '1' an output pulse appears whenever a re-ionizing pulse is applied at A. Erasure is effected by omitting the pulses at A, whereupon all tubes in the memory deionize. It is a limitation of this circuit that tubes can be returned to the '0' condition only by using erasure and this occupies at least 100 μsec.

The two-diode circuit shown in Fig. 3.19 (*b*) overcomes this limitation. Pulses applied to A exceed V_{IG}, so that one tube or the other is always ionized. Suppose that V_0 is ionized to indicate a '0'. A negative-going 'write' pulse applied to the cathode of V_1 will ionize V_1 and reduce the potential of point B to near ground potential. V_0 thus begins to deionize, and at the end of the 'write' pulse V_1 remains the more heavily ionized and thus indicates a '1'. Because the circuit is symmetrical, a similar process allows the memory to be returned to the '0' condition with equal rapidity, either transition being effected in 5 μsec.

Diode Gates

There are several possibilities for the use of diodes as gating elements for either d.c. or a.c. signals.

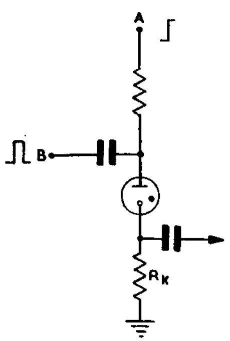


Fig. 3.20. Diode gate for speech and other a.c. signals.

Fig. 3.20 represents a gate useful for a.c. signals. An a.c. input at B will not pass through the tube if the d.c. potential at A is such that the tube anode never exceeds V_{IG}. (It may be necessary to maintain A positive to prevent firing on negative-going peaks of the input at B.) If A is now raised above V_{IG} the tube fires and, provided the negative-going excursions of the input at B do not exceed the d.c. component of the voltage across R_K, the output from the tube cathode follows the input at B with negligible loss. The output impedance is low, being substantially the sum of the tube impedance and that of the a.c. supply to B. Speech signals can be switched by circuits of this type, although noise may prove objectionable if many such switches are connected in series.

Flood and Warman [10] describe a number of logical gates which may be provided by diodes used singly or in combination. Many of these

gates employ the 'pulse-plus-bias' technique (p. 72). The circuit described already in connexion with Fig. 3.20 is, in fact, an 'AND' gate of this type. For reliable operation, a neon diode pulse-plus-bias 'AND' gate requires that

$$V_M > V_P \gtrsim V_B \gtrsim V_{IG}/\sqrt{2} \qquad (3.22)$$

where V_P = positive pulse amplitude.

If a positive bias V_C is applied to R_K instead of at A, a 'NOT' gate results, the tube conducting if a pulse V_P ($>V_{IG}$) is applied at B, but not if a bias V_C ($>V_P - V_{IG}$) is applied to the cathode. For a 'NOT' gate the optimum dimensions of pulse and bias are given by

$$V_P \gtrsim V_{IG}\sqrt{2} \gtrsim 2V_C \qquad (3.23)$$

These two arrangements may be combined to provide a gate with the logical function 'A and B, not C'. For this gate,

$$V_M > V_P \gtrsim V_B \gtrsim V_C \gtrsim V_{IG}/\sqrt{2} \qquad (3.22a)$$

The simple 'OR' gate of Fig. 3.21 (*a*) may also be elaborated to meet more complex requirements. The addition of C and R as indicated by the dotted connexion provides the function 'A and B, or Y', the 'A and B' part of the logic being obtained by pulse-plus-bias methods.

In all the gates described so far a tube conducts to deliver an output, and no tube conducts when an output is not to be delivered. Hendrix [13] has discussed at length the design considerations for 'OR' and 'AND' gates according to the circuits of Fig. 3.21 (*a*) and (*b*) respectively, but based on a different concept. By arranging that one tube or another is always conducting, he maintains the output impedance low and substantially constant.

Fig. 3.21 (*a*) shows an 'OR' gate which has two or more inputs, X, Y, etc.

Each input, X, Y, has two alternative levels: a value V_{XO}, V_{YO}, in excess of V_{IG} and corresponding to a '0' condition and a value V_{X1}, V_{Y1}, higher by about V_M and corresponding to the '1' condition.

If input X is in the '1' condition V_1 will be conducting and the output V_O will be given by

$$V_{01} = (V_{X1} - V_M) \qquad (3.24)$$

If input 'Y' also rises to the '1' condition the voltage across V_2 becomes the same as that across V_1. This voltage, V_M, is insufficient to strike V_2 and the output remains unchanged. If V_X now falls to the '0' condition, however, V_0 falls momentarily until the voltage V_{IG} appears across V_2. At this instant V_2 fires and the output returns to the value

$(V_{Y1} - V_M)$ $[\approx(V_{X1} - V_M)]$. The transient fall in V_0 which accompanies the changeover from V_1 conducting to V_2 conducting (or vice versa) cannot be filtered out simply by connecting a capacitance in shunt with R_L. This would only stretch the pulse beyond the normal duration of 5 μsec.

When both X and Y are in the '0' condition, one of the tubes will

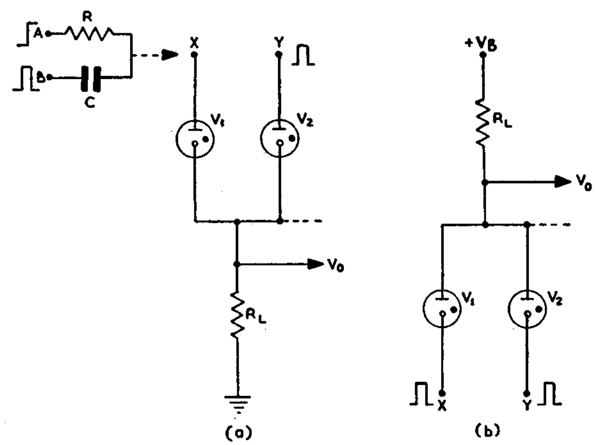


Fig. 3.21. Neon diode gates, (*a*) an 'OR' gate, one input to which is a pulse-plus-bias 'AND' gate, and (*b*) an 'AND' gate.

still be conducting because $V_{X0} \approx V_{Y0} > V_{IG}$. The output V_0 in this case is given by

$$(V_{X0} - V_M) \approx V_{00} \approx (V_{Y0} - V_M) \qquad (3.25)$$

Although it would be possible to operate an 'OR' gate with $V_{X0} \approx V_{Y0} < V_M < V_{IG}$, this would mean that in the condition V_{X0}, V_{Y0}, neither V_1 nor V_2 would be conducting. The output impedance in this state would then rise to the value of the load resistance R_L. By operating the gate in the manner previously described, however, one tube is always conducting and the output resistance R_0 is substantially constant. Thus in all states, R_0 is given by

$$R_0 \approx \frac{(R_X + R_I) \,.\, R_L}{R_X + R_I + R_L} \qquad (3.26)$$

where R_X is the source resistance of input X;
R_Y ($\approx R_X$) is the source resistance of input Y;
R_I is the incremental resistance of the type of tube used for V_1 and V_2.

The 'AND' gate shown in Fig. 3.21 (*b*) operates in a very similar manner to the 'OR' gate. Its output resistance is again given by Equa-

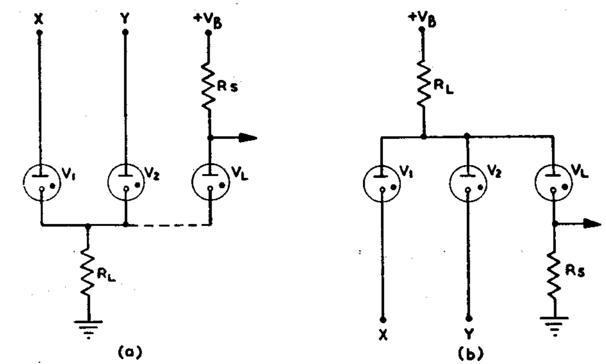


Fig. 3.22. Use of diode V_L as level shifter in (*a*) 'OR' and (*b*) 'AND' gates.

tion (3.25). The output voltage of the 'AND' gate is, however, greater than the lowest input by a voltage V_M, whereas that of the 'OR' gate is lower than the greatest input by a voltage V_M. If one of these gates is required to give an output with the same d.c. levels as the inputs a further diode of the same type may be added as shown in Figs. 3.22 (*a*)

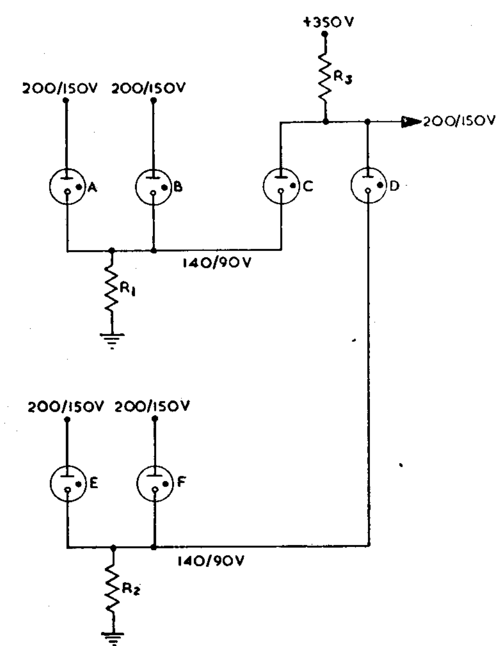


Fig. 3.23. Example of combination of 'OR' and 'AND' gates.

and (*b*). The additional diode V_L acts as a level shifter, raising or lowering the output level by a voltage V_M. The value of R_L must usually be halved, since it now carries the current flowing through V_L in addition to that through one of the other diodes. In general, $R_L \gg R_I$ and $R_S \gg R_I$. It then follows that the output resistance R_0' using a level shifter is greater than that without a shifter and is given by

$$R_0' \approx R_X + 2R_I \qquad (3.27)$$

In some cases it may be required to feed an 'OR' gate into an 'AND' gate, or vice versa. No special steps are then required to effect level shifting, since the two types of gate produce shifts of level which are equal but opposite in direction. This is evident from Fig. 3.23, which shows two 'OR' gates, AB and EF, driving an 'AND' gate, CD. The inputs to the 'OR' gates are at + 200 V in the 'on' condition and + 150 V in the 'off' condition. Assuming a value of 60 V for V_M, the outputs of the 'OR' gates are at 140 V or 90 V, depending on their state. Similarly, the output of the 'AND' gate is at one of two levels 60 V higher than the alternative levels of its input.

EXAMPLE 3.4 Gas Diode Logic Gate

The design of a diode gate depends on the number and nature of other gates to which it is connected. In view of the endless possibilities for such interconnexion, design cannot be reduced to a formal pattern, and each case must be studied from first principles. The following treatment of the circuit shown in Fig. 3.23 is representative. It is based on the paper by Hendrix [13].

The output stage is considered first. Since it operates with the lowest tube currents, these are made as low as is permissible. In this way the operating current of all other tubes is reduced and it is possible to cascade the maximum number of stages before finding the input stage exceeds the maximum permissible tube current.

In gate CD one tube or the other is conducting at any time, but never both. The tube current is clearly lowest when the current through R_3 is at its lowest value, i.e. the output is at its higher value, +200 V. Equating the current through R_3 to $I_{K(min)}$, then Hendrix obtains for the NE2,

$$R_3 = \frac{350 - 200}{I_{K(min)}} = \frac{150}{0{\cdot}2} = \mathbf{750\ k\Omega}$$

When the output is at +150 V the current through R_3 is given by

$$I_{CD(max)} = \frac{350 - 150}{750} \approx \mathbf{0{\cdot}27\ mA}$$

This is the maximum value of tube current in the output stage.

For the gate AB the minimum tube current is again made 0·2 mA. This will correspond to an input at +150 V and an output of 90 V across R_1. In this condition, however, the output from CD is low and the maximum current of 0·27 mA flows through tube C and R_1. Thus the total current through R_1 will be 0·47 mA and R_1 is given by

$$R_1 = \frac{90}{0{\cdot}47} \approx \mathbf{200\ k\Omega}$$

If the output from gate EF is also at +90 V it can happen that tube D, not C, will be conducting. In this case tube A or tube B will carry 0·47 mA so long as both inputs to gate AB are at +150 V. When either input to AB rises to +200 V the voltage across R_1 will rise to 140 V. The current through the conducting tube, A or B, will then reach its maximum value, given by

$$I_{AB(max)} = \frac{140}{200} = \mathbf{0{\cdot}7\ mA}$$

Since the gates AB and EF are identical, $R_2 = R_1 = 200\ k\Omega$ and the tube currents in EF also range from 0·2 to 0·7 mA.

If gates AB or EF are preceded by further gates, the foregoing design procedure is continued. Ultimately a limit is reached at which further stages using similar tubes cannot be cascaded directly without exceeding the tube current ratings. When this arises, tubes of a higher maximum current rating may be used for the early stages. Alternatively, a cathode follower or emitter follower may be used to buffer adjacent gates so that a gate working at comparatively low current can control a subsequent gate operating at higher currents.

Limitations of Diode Gates

Hendrix [13] has shown that it is possible for diode gates using NE2 tubes to operate reliably at rates up to 30 kp/s. He observes that the spread of characteristics between new NE2 diodes is as much as ±25%. On ageing, the maintaining voltage increases fairly abruptly after about 1,000 mA-hours and thereafter assumes an increased value with a spread of only about ±2%. This appears to be due to the sputtering-off of the low work-function oxide coating on the electrodes of new tubes. Thus the higher maintaining voltage corresponds to a tube operating with pure metal electrodes. Hendrix notes the occasional firing delays experienced with pre-aged tubes. He attributes it to the virtual exclusion of light by the deposition of sputtered material on the inside of the envelope. It seems more probable that the explanation lies in the change in photosensitivity of the cathode on losing its low work-function coating. A new cathode will emit photo-electrons when irradiated by visible light. After ageing, however, its sensitivity will be confined to ultra-violet radiation.

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CHAPTER FOUR

Trigger Tubes

Although the cold cathode diode is sometimes used as a relay device, for the majority of applications it suffers from one or more of the following shortcomings:

(1) the breakdown potential is not a controlled characteristic;
(2) striking may be subject to statistical delays;
(3) input and output share the same pair of terminals.

Several kinds of three- and four-electrode relay tubes are now available which eliminate some or all of these shortcomings. In construction, they vary considerably, but in principle they represent a simple logical development from the neon diode.

The main anode–cathode gap of a relay tube has all the characteristics of a gas-filled diode. If the applied voltage exceeds a given value, V_{IG}, breakdown occurs. As long as the current is restricted to the range corresponding to a normal glow discharge, the maintaining voltage, V_M, is substantially constant. As in a simple diode, breakdown cannot occur unless free electrons or ions are present in the gap. Photo-emission or ionization by natural radiation or a radioactive isotope in the envelope produce ion currents so small that the breakdown voltage is not usually dependent on their magnitude. As a rule, only direct sunlight will produce sufficient photo-emission to reduce the value of V_{IG}. In a cold cathode triode, however, the additional electrode provides an alternative means whereby a substantial ionization of the main gap may be produced. When a sufficiently high potential is applied between this trigger electrode and cathode, a discharge occurs between these two electrodes. Provided the trigger current exceeds a critical value known as the *transfer current*, sufficient ionization of the main gap is produced to promote a discharge from anode to cathode. This occurs even though the anode voltage is less than V_{IG}, but it must, of course, be greater than the maintaining voltage, V_M, to which the anode voltage falls once ignition occurs.

The trigger gap also has the characteristics of a cold cathode diode, and accordingly, it, too, cannot strike without a source of primary

electrons. In a relay tube it is usually required that triggering shall occur as quickly as possible once the appropriate potentials have been applied. Accordingly, natural radiation cannot be relied upon to produce primary ionization, as considerable delays would frequently arise between application of the trigger potential and ionization of the gas filling by a cosmic ray. Even the inclusion of a radioactive material, such as uranium oxide or tritium, leaves a significant statistical delay. Accordingly, two alternative types of trigger tube have been evolved:

(*a*) Those in which the cathode is coated with a material of low work function, e.g. an oxide of barium or potassium. Such a coating will emit electrons under the influence of light entering the glass envelope. These electrons then provide the primary ionization necessary for triggering to occur.

(*b*) Tubes using pure metal cathodes, usually molybdenum. The pure metal has a higher work function than an oxide coating, and as a result its photo-emission is confined to ultra-violet wavelengths in the band absorbed by the glass of the tube envelope. Consequently, photo-emission due to external radiation cannot be used to produce primary ionization. An alternative source is provided in a small continuous priming discharge across an auxiliary electrode gap. An external series resistor is used to restrict this discharge to some tens of microamperes, and this low current is not itself enough to cause triggering of other inter-electrode discharges. Some ions or photons do, however, enter the trigger–cathode gap, and so provide primary ionization allowing triggering to occur as soon as the trigger–cathode breakdown voltage, V_T, is reached.

One sees that the relay tube is most aptly named. Application of the requisite trigger potential allows the photo-emission or priming discharge to initiate breakdown of the trigger–cathode gap. Discharge of the trigger circuit in turn supplies the greater degree of primary ionization needed to cause ignition of the main gap at an anode voltage below V_{IG}. Ionization is thus relayed from primary source to main gap once the trigger potential reaches V_T. This justification of the name *relay tube* is only incidental: its true origin lies in the considerable power gain between input and an 'off–on' output.

Ignition Characteristics

Fig. 4.1 represents the ignition characteristic of a typical trigger tube. Provided the voltage applied to the trigger electrode and anode are represented by a point lying within this characteristic, ignition does not

occur. If one or both of these voltages is increased until the working point reaches or passes outside the characteristic, breakdown occurs. Operation is usually confined to the first quadrant, and breakdown is then initially between trigger and cathode or anode and cathode, depending on whether it is the trigger voltage or the anode voltage which is increased to move the working point beyond the characteristic. It will

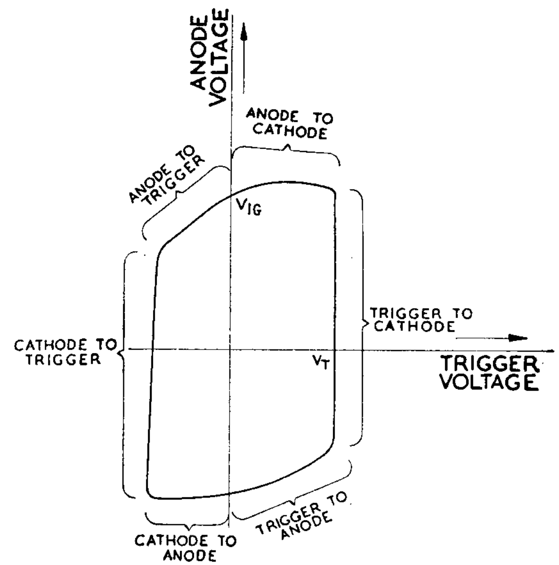


FIG. 4.1. Ignition characteristic of typical trigger tube, indicating electrodes between which initial breakdown occurs.

be seen, however, that if the trigger and/or anode voltage can become sufficiently negative the initial breakdown will occur in either direction between any pair of electrodes. With many circuits, particularly those operating with a.c. supplies, such triggering can occur unintentionally and must be guarded against. If current flows *into* the nominal anode or trigger some of the nickel may be sputtered on to the normal cathode, thereby contaminating the cathode surface. In tubes having pure metal cathodes this produces a permanent change in the otherwise very stable triggering characteristic of the tube.

Once breakdown occurs, the anode voltage falls to the anode-maintaining voltage, V_M. The trigger then acts as a probe in the anode–cathode discharge and tries to assume the trigger maintaining potential, V_N. This it will do provided the trigger current is limited to a few micro-

amperes by a sufficiently large resistance in the trigger circuit. It should be noted that reverse trigger current will flow if the trigger supply potential is allowed to fall below V_N while the anode gap is conducting.

Triggering does not occur instantaneously. Between application of the trigger pulse and the establishment of anode current a delay time T arises which Hough and Ridler [1] express by Equation (4.1).

$$T = t_1 + t_2 + t_3 + t_4 \tag{4.1}$$

where t_1 is the statistical delay in the trigger–cathode breakdown;
t_2 is the formative delay time elapsing between the start of the trigger breakdown avalanche and the formation of the positive space charge;
t_3 is the time required for the trigger discharge to produce breakdown of the main gap; and
t_4 is the time required for the main gap current to rise to a given level.

It has been noted already that the first of these delays, t_1, becomes long and unpredictable if a tube relying on photo-emission is operated in the dark. Even with the inclusion of a small amount of uranium oxide, delays may be as long as 100 msec, and accordingly, high-speed trigger tubes use pure metal electrodes and an auxiliary priming discharge. Provided the priming discharge is sufficiently large (usually between 10 and 250 μA), the statistical variation in t_1 practically disappears. It is then found, however, that t_1 and t_2 decrease if an over-voltage is applied to the trigger–cathode gap before ionization occurs. Over-voltage has the merit of ensuring more reliable operation of a tube despite increases in trigger–cathode breakdown voltage and transfer current during its life.

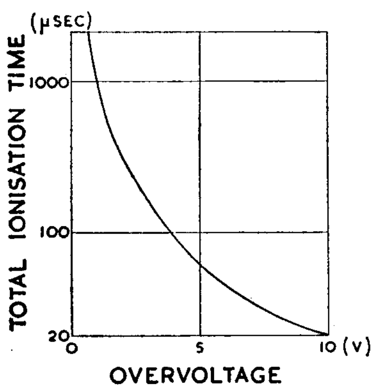


Fig. 4.2. Reduction of ionization time with trigger voltage in excess of critical value.

Over-voltage of the trigger usually produces an increased trigger current after trigger breakdown. This produces a proportional reduction in the transfer time, t_3, for a given anode voltage. The final delay, t_4, is determined largely by the external circuit and supply voltage. The anode current rises more quickly with a high voltage applied through a circuit free of inductance.

Fig. 4.2 summarizes the effect of trigger over-voltage on the total

ionization time, T, in a type CV2434 tube. An increase of over-voltage from 0·5 to 10 V reduces the ionization time from 2 msec to 20 μsec.

A further important factor is the value of anode voltage applied before triggering takes place. From Fig. 4.3 it is seen that raising the

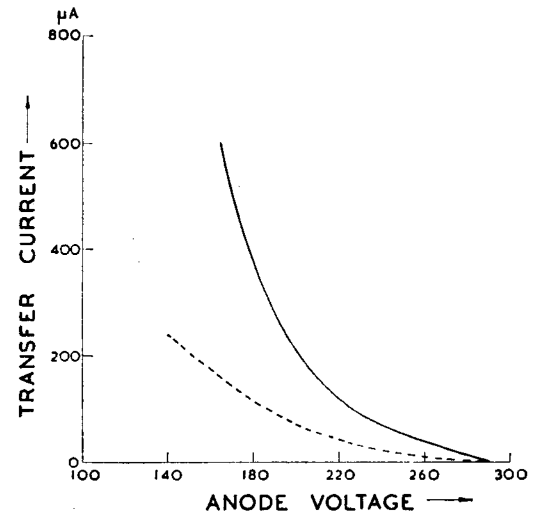


Fig. 4.3. Relation between transfer current and anode voltage for CV2434.

anode voltage above the maintaining voltage rapidly reduces the transfer current. If an excessive anode voltage is applied the transfer current approaches zero, i.e. the anode breakdown voltage, V_{IG}, is reached and breakdown occurs spontaneously. There is, however, a decided advantage in designing for operation with an anode potential of about $0{\cdot}8\,V_{IG}$. With this value, reliable operation is provided with a low transfer current drawn from the trigger.

Shielded-anode Tubes

Whereas the typical molybdenum–cathode trigger tube provides an anode maintaining voltage, $V_M \approx 110$ V and an anode breakdown voltage, $V_{IG} \approx 290$ V, shielded-anode tubes may be operated with anode supply voltages up to about 400 V. As a result, the following advantages are obtained:

(1) For the same value of $I_{K(av)\,(max)}$ (typically ~25 mA), more power may be delivered to the load.
(2) The electrical efficiency of power switching is higher.

(3) Tubes may be operated from rectified a.c. mains supplies of up to 270 V r.m.s. without the power loss or complication of a voltage attentuator.

In construction, the shielded-anode tube differs from the simpler form of trigger tube in having an additional shield electrode between the main anode and cathode. This shield (or 'second anode') controls the potential gradients between anode and cathode prior to breakdown. If it is biased to a suitable proportion (typically 50–75%) of the main anode potential it largely restricts distortion of the anode–cathode potential gradient due to space charge (p. 13). As a result, the point of breakdown is delayed until the potential between an adjacent pair of electrodes reaches a critical value. Typical critical values are $V_T \approx 120$ V, $V_{IG(A1)} \approx 250$ V, $V_{(A1-A2)\ (max)} \approx 180$ V.

Of the shielded-anode tubes currently available, both the Mullard Z806W and the Ericsson GPE/120T provide stable, close-tolerance trigger characteristics with small hysteresis effects (p. 67). The maximum spread in the value of V_T is of the order of only 5% throughout life, including hysteresis.

Two methods of supplying the shield electrode (A2) are commonly used. When the main anode (A1) is connected to a d.c. supply A2 may be returned to any fixed voltage in the appropriate range. Alternatively, if the anode is connected to a supply of rectified, unsmoothed a.c. A2 is returned to a potential divider across this supply. In either case the circuit supplying A2 should have a source impedance of about 100 kΩ, so that, once triggering occurs, the cathode current is principally due to current from A1.

The Elesta ER32 contains a shield anode, the potential of which is set by the gas filling. No external connexion is needed or provided.

In the shielded-anode tube three discharge gaps must be extinguished: trigger–cathode, A1–cathode, and A2–cathode. The techniques available are the same as for other types of tube (pp. 77 *et seq.*), but the A1–cathode and A2–cathode gaps must both be extinguished before working voltage is re-applied to either gap. This requirement is met automatically if A2 is connected to a potential divider between cathode and the supply to A1.

High-speed Trigger Tubes

Shield electrodes are used also in high-speed trigger tubes, the development of which has been described by Hough and Ridler [1]. Most of the foregoing remarks apply also to these tubes, although the shield electrode is provided primarily to assist in rapid deionization. This it

does by providing an electric field which removes ions when the anode–cathode voltage is depressed below the maintaining voltage.

In the S.T. & C. high-speed trigger tube, type G1/371K, the combination of gas-filling and electrode structure provides a deionization time of only 30 μsec. The tube is therefore useful for high-speed counting and logical circuits. For these applications the advantages of short ionization and de-ionization times outweigh the disadvantages of rather high working voltages ($V_T = 177$–192 V, $V_M = 172$–188 V).

In the G1/371K it is also permissible to effect triggering by coincident application of a positive pulse to the shield and a negative pulse to the trigger. Breakdown from shield to trigger is thus produced which leads to breakdown of the main anode–cathode gap. This method of triggering provides useful possibilities in logical circuit design.

Glow Thyratrons

In the various trigger tubes described above a small amount of primary ionization is continuously present in the anode–cathode and trigger–cathode gaps. Trigger–cathode breakdown is produced by increasing the trigger–cathode potential beyond the breakdown potential. Anode–cathode breakdown then results due to the consequent heavy ionization of the anode–cathode gap.

In the Cerberus glow thyratron tubes made in Switzerland a different mode of triggering is employed. Ions produced in a priming discharge are prevented from entering the main anode–cathode gap by a small negative potential (~5 V) applied to a perforated control grid close to the main cathode and between cathode and anode. When the grid bias is removed or made positive sufficient ions pass into the main anode–cathode gap to reduce the breakdown potential below the applied anode voltage. An anode–cathode discharge then occurs which can be extinguished only by reducing the anode–cathode voltage for a time exceeding the deionization time.

The pure molybdenum cathode of the Cerberus GT21 endows it with the stability and long life typical of tubes using pure metal cathodes. The high anode breakdown voltage ($V_{IG} \approx 450$ V) and moderately high mean cathode current (10–40 mA) are attractive features. The small control signal (5 V) is readily provided by transistor circuits at cathode potential.

Characteristics and applications of these tubes have been described by Pun and Mitter [2] and, at greater length, by Henry [3]. As in the more conventional trigger tube, the ionization time of a glow thyratron is reduced by increasing the anode voltage. Unlike the trigger tube, the

glow thyratron ionizes more slowly if the control grid is made more positive than the critical triggering potential. Also the grid of the GT21 acts as a probe in the priming discharge and so tends to carry a small current, even when the tube is in the 'off' condition. With the recommended value of priming current ($\sim$100 μA) flowing between the main cathode and auxiliary cathode, the grid current will not exceed 1 or 2 μA at a bias of -5 V. If the grid is connected to a voltage source of high impedance ($>$1 MΩ) the *IR* drop due to grid current becomes significant.

Once the main discharge is initiated, the grid acts as a probe in the main anode–cathode gap. It may then be regarded as a voltage source of 100 V with an impedance of 100 kΩ.

From the foregoing, it will be understood that it is not convenient to connect the grid to circuits of very high impedance. In practice, variation of grid circuit impedance over the range 50–500 kΩ produces a negligible effect on tube performance.

Deionization and Recovery Time

The comments made in Chapter Three on the deionization time of diodes (p. 40) apply also to trigger tubes. Thus recovery is most rapid

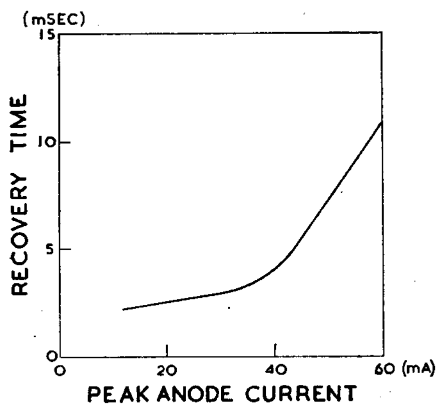


Fig. 4.4. Relation between anode gap recovery time and peak anode current for CV2434.

if anode and trigger potentials fall during recovery to about half their maintaining potentials.

In trigger tubes the picture is complicated by the possibility of breakdown in either trigger or anode gap when normal pre-strike potentials are re-applied. The two gaps may be regarded as having separate

recovery times, distinguishable by the different results obtained, depending whether it is the trigger or the anode potential which is returned to a potential near breakdown after the deionization period. Thus Fig. 4.4 shows how the anode–cathode gap recovery time varies with peak anode current in the CV2434. If an anode current of 40 mA has been established the anode voltage must be held at zero for a

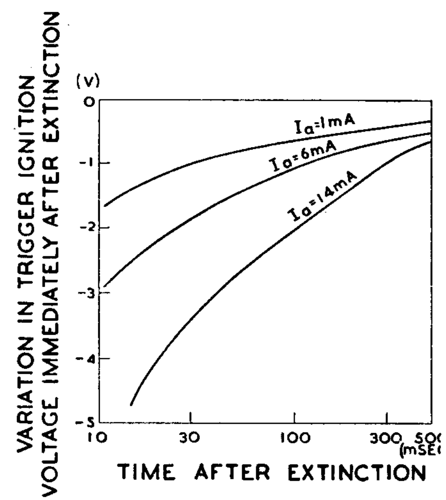


FIG. 4.5. Trigger voltage depression ('hysteresis') as a function of recovery time and anode current.

minimum of 4 msec before it can be returned to 220 V without re-establishing the discharge. This is so even though the trigger potential may be held well below the normal breakdown trigger potential. Fig. 4.5 shows how the trigger-cathode breakdown voltage, V_T, of the same tube is reduced considerably when a large anode current is followed by a brief extinction period. This 'trigger hysteresis' is present in some degree in all trigger tubes. Crowther and Smith [4] have established that it is due to heating inside the tube. For the Z803U, the formative and recovery time-constants are approximately equal and of the order of 8 sec. Tubes are now available which have been designed to reduce trigger hysteresis to 1 or 2% of V_T.

Hough and Ridler [1] have described the processes involved in deionization. The removal of electrons may be effected within a few microseconds by the application of a suitable field. Ions are also removed by an applied field, though rather more slowly on account of their lower

mobility. In many tubes the principal factor determining the deionization time lies in the presence of atoms of gas which have been raised to metastable states of excitation by the discharge. Electrons have not been removed from these atoms, but only raised from one orbit to another. Consequently, these metastable atoms carry no net charge and so cannot be removed by an applied field.

Although metastable atoms may have energy levels as high as 10 V, electrons do not remain in these orbits for very long. In less than 1 μsec they decay to levels of about 3 or 4 V. These lower energy levels exceed the work function of an oxide-coated cathode, but not that of a pure metal cathode. With a coated cathode, therefore, metastable atoms may produce either secondary emission or photo-emission, so long as they remain in the tube. The life time of a metastable atom may be of the order of 1 msec, and accordingly, tubes with oxide-coated cathode have correspondingly long deionization times.

When pure metal cathodes are used the metastable atoms do not possess individually sufficient energy to produce either secondary or photo-emission from the cathode. However, relatively long deionization times may still be obtained because so long as they persist, metastable atoms collide to produce further ionization.

By introducing some hydrogen in the gas-filling it is possible to produce quenching of metastable atoms within a few microseconds. Hough and Ridler describe the application of this technique to the design of high-speed trigger tubes such as the G1/370K and G1/371K. In these tubes deionization times as short as 20 or 30 μsec are obtained. Deionizing agents cannot be used with oxide-coated cathodes, as the cathode surface is rapidly destroyed by such additions to the gas filling.

Current Triggering

When a high-impedance input is applied to the trigger the loading due to the transfer current must be considered. As the trigger potential approaches the critical value, V_T, a small current ($\sim 10^{-10}$ A) flows from trigger to cathode. In the circuit of Fig. 4.6 (*a*) it is therefore a prerequisite for triggering that

$$V_S > V_T \tag{4.2}$$

As the trigger potential reaches V_T the trigger–cathode current increases very rapidly and enters a negative-resistance zone, DE in Fig. 2.2. If the value of R_T is not large enough to stabilize the discharge in this zone the trigger current increases to a value I given by

$$I = \frac{(V_S - V_N)}{R_T} \tag{4.3}$$

For triggering to occur,

$$I \geqslant I_T$$

i.e.

$$I_T \leqslant \frac{(V_S - V_N)}{R_T}$$

This will be satisfied if

$$R_T < \frac{(V_T - V_N)}{I_T} \tag{4.4}$$

Thus triggering is assured if Relations (4.2) and (4.4) are both satisfied for the maximum value of I_T during tube life.

The maximum value of R_T given by Relation (4.4) depends on the tube. Typically it is about 300 kΩ. Provided this limit is respected, V_S need exceed V_T by only a few millivolts, since the pre-strike current produces only a very small voltage drop across R_T. If the limiting value of R_T is exceeded, however, triggering will not occur until Relation (4.5) is satisfied.

$$V_S > V_N + I_T R_T \tag{4.5}$$

In this case the critical value of V_S for a given value of R_T becomes dependent on I_T. Since I_T is likely to vary during the tube life, this is not a mode of operation providing triggering at a consistent level of V_S.

If difficulty is experienced in satisfying Relation (4.4) it should be remembered that I_T may be decreased by an increase in V_A. The maximum value of V_A is determined by the danger of exceeding the anode–cathode breakdown voltage, V_{IG}.

Capacitor Triggering

When the source impedance exceeds the limit set by Relation (4.4), capacitor triggering can be a useful circuit technique. It raises the limiting value of R_T by some three orders of magnitude, though at the expense of a delay which can rise to about 0·1 sec in an extreme case. This is due to the addition of the capacitor, C_T in Fig. 4.6 (*b*), between trigger and cathode.

The addition of C_T allows an avalanche to occur when the discharge enters the negative resistance part of its characteristic. As the trigger potential falls from V_T to V_N, current is drawn from the capacitor, and if the charge so removed is sufficient, triggering occurs. If not, the discharge may stabilize at a current determined by R_T. Alternatively, squegging (relaxation oscillations) may occur, and triggering may occur after several squegging pulses.

Manufacturers usually provide data covering the minimum value of

C_T ensuring triggering with specified anode voltages. A rough estimate of this value, $C_{T(crit)}$, may be arrived at by calculating the capacitance which, in discharging from V_T to V_N, loses a charge equal to the product of the transfer current, I_T (for the appropriate anode voltage), and the ionization time, T_I, without trigger over-voltage. Thus:

$$C_{T(crit)} \approx \frac{I_T \,.\, T_I}{V_T - V_N} \qquad (4.6)$$

In the absence of precise information it is advisable to adopt a value of C_T two or three times larger than that given by Relation (4.6).

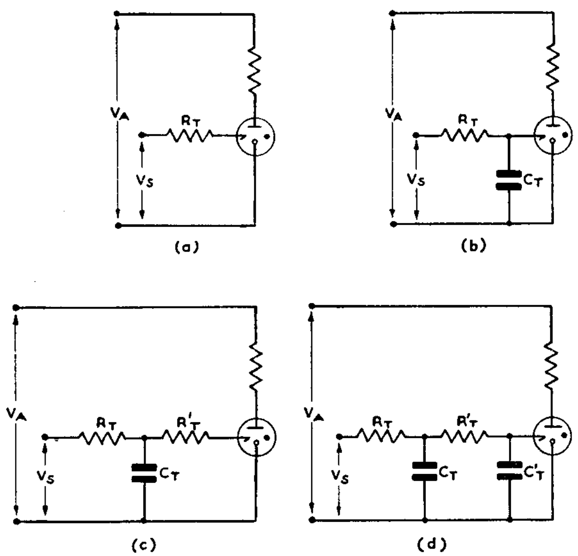


Fig. 4.6. Trigger circuits arranged for (*a*) current triggering, (*b*) capacitor triggering, (*c*) capacitor triggering with R_T to limit trigger current, (*d*) alternative arrangement used where C_T is large.

If $C_T R_T$ is large there is an appreciable delay between the application of V_S and the rise of trigger potential to the critical voltage. This is the basis of various timing circuits discussed later. When a value of C_T is used much exceeding that given by Relation (4.6) a series resistor, R_T' (usually a few kilohms), should be used, as in Fig. 4.6 (*c*), to limit the trigger current to a value which will not damage the tube.

When the values of R_T and C_T are both large even Fig. 4.6 (*c*) does not represent the optimum circuit. The use of R_T' limits the range of the negative resistance part of the trigger discharge characteristic over which instability is attained. In consequence, a slightly higher pre-strike current

is drawn before the avalanche occurs. This effect (which can be significant when a vacuum photocell is used in place of R_T), can be overcome by adding C_T' as in Fig. 4.6 (*d*). Here C_T' has the value given by Relation (4.6) and $R_T'C_T' \ll R_T C_T$.

Maximum circuit stability requires that the drop across R_T due to the pre-strike current, I_P, shall be negligibly small compared with V_T. Thus:

$$R_T \ll V_T/I_P \qquad (4.7)$$

With tubes having oxide-coated cathodes the value of I_P depends on the level of ambient illumination. For such tubes, Relation (4.7) cannot be used to set an upper limit to the value of R_T.

Pulse Train Triggering

In Figs. 4.7 (*a*) and (*b*) circuits are shown which closely resemble those of Figs, 4.6 (*c*) and (*d*). In Fig. 7 (*a*), however, the capacitor C_P is connected to a source of positive-going pulses. If V_S is made positive, but

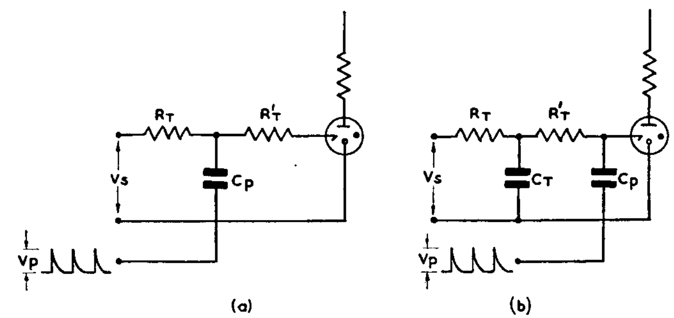


Fig. 4.7. Pulse train triggering used (*a*) with high impedance input to trigger, (*b*) in timing circuit.

sufficiently less than V_T, no trigger current will flow during the intervals between pulses. At each pulse, however, the trigger potential is momentarily raised, and if the sum of V_S and the positive pulse excursion, V_P, exceeds V_T capacitor triggering occurs.

In timing circuits long but accurate delays can be obtained with reasonable values of C_T only if R_T can be made large. This can be done if the circuit of Fig. 4.7 (*b*) is used to reduce the mean pre-strike current to a low value. Between positive-going pulses little or no pre-strike trigger current flows. Relative to the pre-strike current during the positive pulses, the mean trigger current is thus reduced in proportion to the space/mark ratio of the pulse train. Due to the relation between

trigger over-voltage and time, however, any attempt to reduce mean trigger current by using extremely narrow pulses is largely ineffective. The shorter pulses require a larger pre-strike current, and there is little or no saving once the pulses are reduced to 1 msec duration. Large space/mark ratios become beneficial only with pulse repetition frequencies of 100 c/s or less. The advantages of capacitor triggering remain, of course.

When first described [5], pulse train triggering provided a valuable technique whereby large timing resistances or other high-impedance sources could be used with the tubes then available. Since that time, trigger tubes have been developed [6] in which the voltage/current characteristic of the trigger–cathode gap shows a negligible region of corona stabilization. With these tubes, adequate performance can usually be obtained using capacitor triggering. The Z803U, for example, provides reliable triggering with values of R_T as high as $10^9\ \Omega$. If higher values of R_T were used, variations of leakage resistance would too greatly influence performance.

Pulse train triggering depends on both the d.c. component, V_C, of the trigger voltage and also on the positive pulse amplitude, V_P. In fact,

$$V_{C(\text{crit})} + V_P = V_T + V_O \tag{4.8}$$

where V_O is the trigger over-voltage required by the limited width of the positive pulse. The value of $V_{C(\text{crit})}$ will thus vary in sympathy with changes in V_P. This feature has been employed by Hercock and Neale [7] and by Young [8] in the design of timing circuits providing intervals bearing a power-law relation to the voltage of the a.c. mains supply. The design of such circuits is discussed at pp. 118 *et seq.* Other relations are undoubtedly possible.

Pulse-plus-bias Triggering

A single positive pulse applied through C_P in Fig. 4.7 (*a*) will produce triggering if a positive bias, V_S, is applied exceeding the value of $V_{C(\text{crit})}$ given by Equation (4.8). In tubes with pure metal electrodes the high stability of the trigger potential makes it practicable to set the bias V_S to a value only 2 or 3 V below V_T. A pulse of a few volts amplitude will then produce triggering. Depending on the value of V_S chosen, the pre-strike current is reduced to less than 1% of that with capacitor triggering: smaller values of V_S yield even larger factors of reduction.

By adjusting the value of bias voltage applied, it is possible to set the sensitivity so that triggering occurs at any chosen pulse amplitude between $0{\cdot}02V_T$ and V_T. To obtain high stability of the trigger potential, however, it is essential to ensure that reverse trigger current does not flow. This generally dictates that $V_S > V_N$. The critical pulse amplitude, $V_{P(\text{crit})}$, is then given by

$$0{\cdot}02V_T < V_{P(\text{crit})} < (V_T - V_N) \tag{4.9}$$

Because the trigger gap has an increased sensitivity after a large anode current has passed (p. 67), low values of $V_{P(\text{crit})}$ can be realized with stability only if a recovery period of 1 sec or more is allowed between extinguishing the discharge and re-application of the full d.c. bias.

When the bias is applied to R_T (Fig. 4.7 (*a*)) the trigger rises to the bias potential with a time-constant C_PR_T. If pulse and bias are individually insufficient to produce triggering short pulses can produce triggering only if the bias anticipates them by more than a minimum time interval. Pulse-plus-bias triggering is thus better described as 'bias-before-pulse' triggering. It is essentially a logical process of 'A before B' rather than a true 'AND' gate.

The rise of trigger potential to the critical bias, $V_{C(\text{crit})}$, occurs in a time, t, given by

$$t = C_PR_T \log_e \frac{V_S}{V_S - V_{C(\text{crit})}} \tag{4.10}$$

Knowing the amplitude of the pulse, V_P', *at the trigger*, Equations (4.8) and (4.10) may be combined to deduce the minimum time by which the bias must anticipate the pulse.

$$t_{\min} = C_PR_T \log_e \frac{V_S}{V_S - (V_T + V_O + V_P')} \tag{4.11}$$

Attempts to reduce $t_{\min}$ by reducing C_PR_T lead to differentiation of the applied pulse waveform. The pulse appearing at the trigger may then hold the trigger above V_T for too short a time to ensure ionization.

If the input pulse is of rectangular form differentiation by C_PR_T causes its trailing edge to apply a negative-going pulse to the trigger. In the absence of d.c. bias this may momentarily increase the anode–trigger voltage sufficiently to produce spurious triggering due to anode–trigger breakdown. This danger is removed if the applied anode voltage is made smaller than the anode breakdown voltage by a voltage at least equal to the input pulse amplitude.

Often the applied pulse rises exponentially. Because C_PR_T forms a differentiating network, the pulse V_P' at the trigger is then of smaller amplitude than the applied pulse V_P. This loss is, of course, most serious when the time-constant C_PR_T is made small in an attempt to reduce $t_{\min}$. Flood and Warman [9] have studied the effect and conclude that

the peak pulse amplitude V_P' at the trigger is related to the input pulse amplitude V_P by Equation (4.12).

$$V_P' = y \,.\, V_P = (e^{-\alpha x} - e^{-x})\,(1 - \alpha)\,.\, V_P \qquad (4.12)$$

where $\qquad x = t/C_P R_T$

and $\qquad \alpha = \dfrac{\text{time-constant of applied pulse}}{C_P R_T}$

From Fig. 4.8 it will be seen that if $\alpha = 1$, V_P' is only 37% of the input pulse amplitude. If the trigger is to receive 85% of the input pulse amplitude, $C_P R_T$ must be at least 20 times the input pulse time-constant.

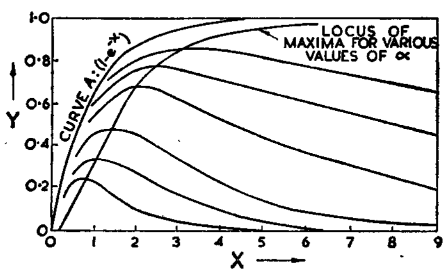


Fig. 4.8. Proportion, y, of exponentially rising input pulse reaching trigger through CR coupling of $1/\alpha$ times the pulse time-constant. Abscissa, x = time ÷ coupling time-constant.

In making $C_P R_T$ large enough to satisfy these requirements, it may be necessary to use a rather large value of C_P. There is then some danger that when trigger–cathode breakdown occurs, a trigger current will flow which exceeds the permitted maximum, $I_{T(max)}$. A series resistor, R_T', must therefore be used in the trigger circuit having a minimum value given by Equation (4.13).

$$R_T'{}_{(min)} = \frac{(V_S + V_P - V_N)}{I_{T(max)}} \qquad (4.13)$$

When pulses are to be applied to several tubes in parallel it may be desirable to use a higher value of R_T' so that after a tube fires it does not unduly load the pulse generator.

Design procedures for pulse-plus-bias triggering will be given when ring counters are discussed. It will then be shown that it is often necessary to add a reference line and catching diodes to define accurately the bias potential applied to each tube.

Voltage Transfer Triggering

By connecting the trigger electrode to the output of an 'AND' gate as in Fig. 4.9, a trigger circuit is obtained differing from the pulse-plus-bias in these respects:

(1) A true 'AND' logic is obtained (not 'A before B').

(2) After triggering, a momentary interruption of the h.t. supply is followed by restriking. (With pulse-plus-bias triggering, the tube does not restrike unless a further pulse is applied.)

(3) Bias stabilization is not necessary.

(4) Signal reshaping is not necessary.

Against these advantages, it is found that counting circuits become rather clumsy and tubes are required having higher anode breakdown voltages than for pulse-plus-bias counters.

Voltage transfer triggering is not a distinct method of triggering (such as current triggering or capacitor triggering) so much as a method of transferring signals from one tube to another. The triggering process usually corresponds to current triggering, though capacitor triggering can occur. In the design of the logic gates care must be taken to see that the self-capacitance of the diodes does not cause pulse-plus-bias triggering to occur before the functional (e.g. 'AND') requirements of the gate are satisfied. In Fig. 4.9, for example, the sudden appearance of an input at B might cause a momentary positive pulse to appear on the trigger even though A is still in the 'off' condition.

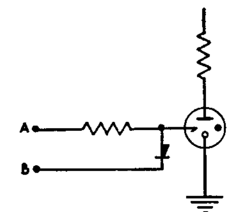


Fig. 4.9. Voltage transfer triggering.

Beesley [10] has described problems arising in the use of voltage-transfer methods for a wide range of operations. Not only 'AND' gates are used in these circuits: there is no restriction on the type of gate which may precede a tube.

Cathode Triggering

All the triggering methods discussed above have involved raising the trigger potential. A less often used alternative comprises depressing the cathode potential until the necessary trigger–cathode voltage is produced.

In Fig. 4.10 (*a*) V_2 is triggered by a negative-going pulse applied to its cathode from the anode of V_1 via C_1. This provides a means of striking

V_2 sometimes useful for its relative freedom from interaction with circuits attached to the trigger of V_1. It has the disadvantage, however, that, to prevent the flow of reverse trigger current, R_K must be kept low:

$$\frac{R_K}{(R_A + R_K)} \cdot (V_B - V_M) \leqslant (V_S - V_N)$$

whence
$$R_K \leqslant R_A \Big/ \left[\frac{V(_B - V_M)}{(V_S - V_N)} - 1\right] \tag{4.14}$$

Consideration of the term $(V_B - V_M)/(V_S - V_N)$ shows that, due to tolerances on the values of V_M and V_N, reasonably large values of

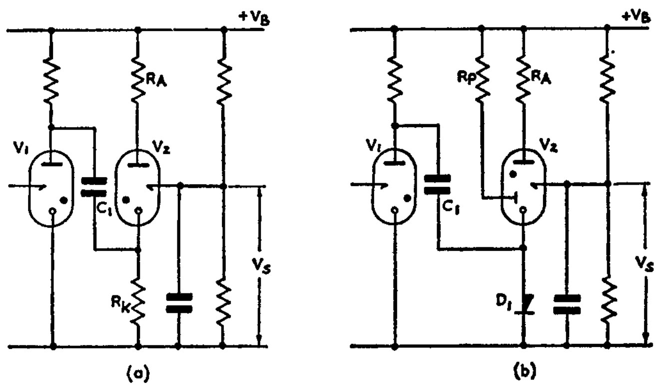


Fig. 4.10. Triggering by negative pulse from V_1 to cathode of V_2, (*a*) with resistance, and (*b*) with diode in cathode of V_2.

R_K may be used only by using a low value of V_B and by making V_S approach $V_{T(min)}$.

The pulse required to trigger V_2 must have an amplitude exceeding $(V_{T(max)} - V_S)$. Cathode triggering therefore requires a low-impedance source if the input is applied to a cathode resistor as in Fig. 4.10 (*a*). The discharge of C_1 through V_1 and R_K provides such a source.

Where rapid continuous triggering of V_2 is required (e.g. on each cycle of an a.c. power supply), the large value of C_1 dictated by the circuit of Fig. 4.10 (*a*) can lead to an excessive mean cathode current in V_1. The circuit shown in Fig. 4.10 (*b*) allows a relatively high-impedance trigger source to be used, even though R_A may be quite low. Before either tube is triggered, D_1 is held conducting by the small priming current flowing through V_2. (In the absence of a priming anode, 10 or 20 MΩ may be connected between V_2 cathode and the h.t. rail.) When V_1 fires, D_1 is

cut off and the cathode of V_2 is driven negative to a voltage $-(V_B - V_M)$. The peak inverse rating of D_1 must be sufficient to withstand this voltage, since this will be applied (decaying with a time-constant C_1R_P) until V_2 strikes. Thereafter the cathode current of V_2 causes D_1 to conduct, and the cathode of V_2 is thereby held at ground potential. With this arrangement there is no danger of reverse trigger current provided $V_S > V_{N(max)}$.

Extinguishing Trigger Tubes

The methods by which a trigger tube may be extinguished are similar to those available for the diode. In trigger tubes, however, the anode–cathode and trigger–cathode discharges must both be extinguished and in the correct sequence. With most tubes – particularly those having pure metal cathodes – it is important to ensure that reverse trigger current flows at no time. This means the anode gap must be extinguished before the trigger gap, as otherwise the trigger will behave as an auxiliary cathode. If this happens, traces of the trigger electrode material will be sputtered on to the cathode, contaminating its surface and rendering the tube characteristics unstable.

Series-switched Extinguishing

The simplest way of extinguishing a cold cathode tube comprises opening a switch in the h.t. supply. With a close-tolerance tube the switch must remove the trigger bias *after* the anode discharge has collapsed. In Fig. 4.11 (*a*) this may not happen, since the anode decoupling capacitor may delay extinction of the anode circuit longer

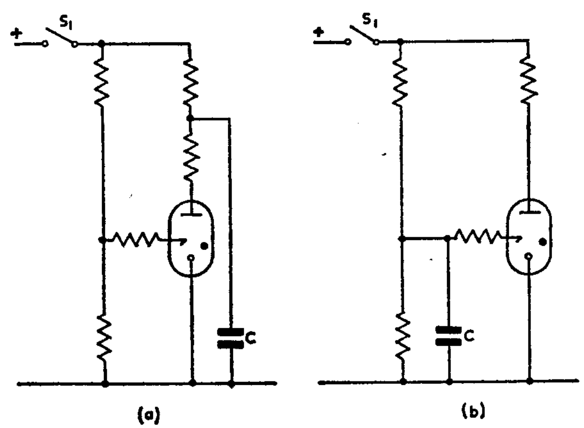


Fig. 4.11. Series-switched extinguishing circuits (*a*) unsuitable, and (*b*) suitable for close-tolerance tubes.

than it delays extinction of the trigger circuit. Fortunately Fig. 4.11 (*b*) represents a more commonly met arrangement, and here the capacitor C ensures the correct sequence of quenching.

Shunt-circuit Extinguishing

A common requirement, notably in timing and control circuits, is that the trigger tube shall energize a relay which then operates to extinguish the tube. Fig. 4.12 (*a*) shows one such arrangement. When V_1 fires,

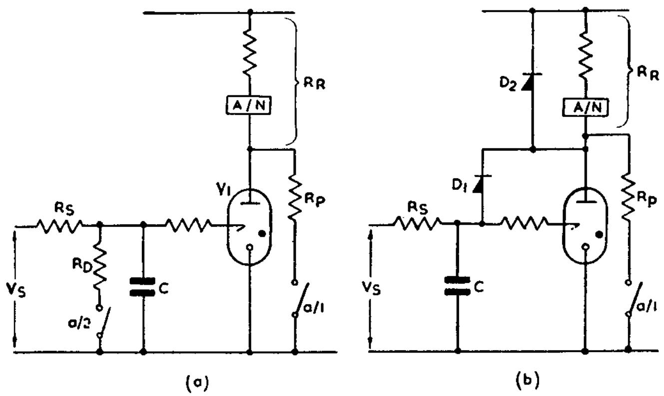


Fig. 4.12. Shunt-circuit extinguishing arrangements. In (*a*) failure of contacts $a/2$ may lead to damage to tube. In (*b*) no such danger arises.

trigger and anode fall to their maintaining voltages, V_N and V_M respectively. Relay A/N energizes and contacts $a/1$ close the latching circuit through R_P. The anode discharge is then extinguished if the potentiometer formed by R_P and the relay resistance, R_R, pulls the anode potential below V_M, i.e. if

$$R_P < \frac{V_M}{(V_B - V_M)} \cdot R_R \tag{4.15}$$

Relay A/N also closes contacts $a/2$ so that, by a similar process, the trigger discharge is extinguished if

$$R_D < \frac{V_N}{(V_S - V_N)} \cdot R_S \tag{4.16}$$

To allow for component and supply voltage tolerances, it is advisable not to exceed 0·7 of the value for R_P and R_D given by Relations (4.15) and (4.16).

The arrangement shown in Fig. 4.12 can cause reverse trigger current

to flow unless $a/2$ is made to close after $a/1$. Even then, a single failure of $a/1$ (due to dust or contamination of the contacts) will cause permanent damage to the tube.

These limitations are removed in the circuit shown in Fig. 4.12 (*b*). Before operation of the contacts $a/1$, the anode is above the trigger potential, and D_1 is therefore cut off. When the relay contacts $a/1$ close, however, the hold-on resistor R_P pulls the anode potential not only below V_M (thereby extinguishing the anode discharge) but also below the trigger-maintaining potential, V_N. D_1 then conducts to lower the trigger potential and extinguish the trigger discharge. Correct operation requires that

$$R_P < \frac{V_N}{(V_B - V_M)} \cdot R_R \tag{4.17}$$

The back resistance of D_1 must be high compared with R_S: in timing circuits a good silicon diode may be required. In timing circuits it may also be convenient to put $R_P = 0$ so that D_1 will discharge C completely ready for the initiation of another timing interval.

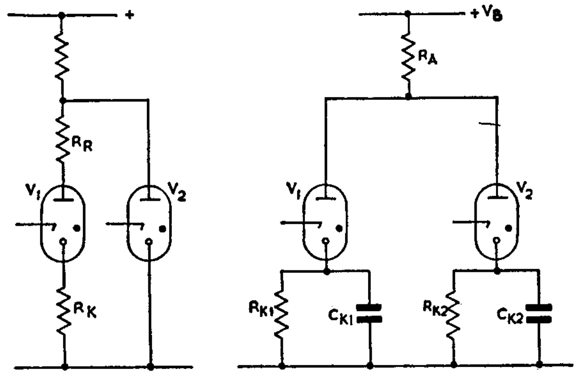
The diode D_2 is needed because the reset contacts, S_1, have been put in the shunt circuit path. Opening S_1 interrupts the current through the relay. An inductive kick is thereby produced which, but for D_2, would swing the anode potential of V_1 above V_B. If the upward swing took the potential above V_{IG} anode–cathode breakdown would occur. D_2 catches the anode potential at V_B and provides a discharge path for the energy stored in the relay coil. The peak inverse voltage applied to D_2 is $V_B . R_R/(R_P + R_R)$ and the peak current is $V_B/(R_P + R_R)$. It is good practice to use a surge-absorbing diode across any inductive anode load, even when the reset contacts are in the h.t. line as in Fig. 4.12 (*a*).

Common-anode-load Extinguishing

A special form of shunt circuit extinguishing is that in which two or more tubes share a common anode load as in Fig. 4.13 (*a*). A cathode load R_K or a relay R_R is energized so long as V_1 is conducting. When V_2 is subsequently triggered the anode discharge of V_1 is extinguished, provided that the anode-maintaining voltage of V_2 is less than that of V_1. If this is not the case extinction can still be ensured by returning the cathode of V_2 to a potential sufficiently negative with respect to the cathode of V_1.

Often it is inconvenient to have to grade the anode-maintaining voltages or the cathode potentials of the tubes in this way. Alternative methods involve shunting R_R or R_K by a suitably large capacitance.

The more common arrangement comprises shunting the cathode resistors, as shown in Fig. 4.13 (*b*). While V_1 is conducting, C_{K1} charges to an equilibrium value. When V_2 is triggered the cathode of V_1 is held positive by C_{K1}, while that of V_2 is initially at ground because C_{K2} has yet to charge up. If the circuit time-constants are correctly chosen, V_1 cathode will remain above V_2 cathode long enough for V_1 to deionize.



(a) (b)

Fig. 4.13. Common-anode-load extinguishing (*a*) with V_2 chosen to have a lower maintaining voltage than V_1, and (*b*) with similar tubes.

Flood and Warman [9] have shown that a conservative estimate of the minimum value of C_K providing reliable quenching may readily be calculated.

If in Fig. 4.13 (*b*) V_1 is being extinguished its cathode potential, V_{K1}, is given by

$$V_{K1} = V_K \,.\, e^{-\alpha t} \qquad (4.18)$$

where

$$\alpha = 1/R_K C_K \qquad (4.19)$$

and time, t, is measured from the instant at which V_2 is struck.

The cathode potential, V_{K2}, of V_2 is given by

$$V_{K2} = V_K \,.\, (1 - e^{-\beta t}) \qquad (4.20)$$

where

$$\beta = \alpha \,.\, \left(1 + \frac{R_K}{R_A}\right)$$

Now as V_{K1} decays, the voltage across V_1 increases. The danger is that the tube may restrike when the applied voltage rises to V_M. Since V_2 is conducting, this state is reached when the voltages across V_1 and V_2 are equal, i.e. when $V_{K1} = V_{K2}$. Provided V_1 has deionized before this

state is reached, there is no danger of its restriking. Since the tube deionization time, T_D, is normally quoted with reference to a voltage substantially in excess of V_M, it is sufficient to choose C_K so that $V_{K1} = V_{K2}$ at a time $T = T_D$.

Thus combining Equations (4.18) and (4.20),

$$e^{-\alpha T} = 1 - e^{-(1+R_K/R_A)\,.\,\alpha T} \qquad (4.21)$$

From Equation (4.21), αT has been plotted in Fig. 4.14 as a function of R_K/R_A. Thus, once the ratio R_K/R_A is known, αT may be read from

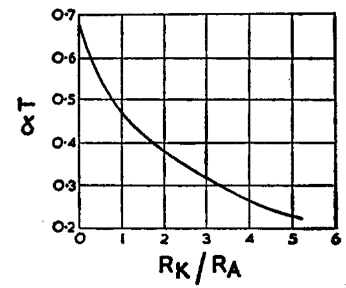


Fig. 4.14. Design curve for determining minimum value of $C_K R_K$ in Fig. 4.13 (*b*). $(C_K R_K)_{min}$ = Deionization time ÷ αT.

the curve of Fig. 4.14 and may be inserted in Equation (4.19a) to derive $C_{K(min)}$.

$$C_{K(min)} = \frac{T_D}{\alpha T \,.\, R_K} \qquad (4.19a)$$

Design Procedure for Cathode Capacitor Extinguishing

(*a*) Choose a suitable value for V_B. There are two possibilities:

(i) That (any) one tube must strike when V_B is applied. In this case, choose V_B so that

$$V_B > V_{IG(max)}$$

(Unless the rate of rise of h.t. voltage is restricted, more than one tube may strike simultaneously.)

More commonly, (ii) no tube must fire due to anode–cathode breakdown. Thus V_B is chosen so that

$$V_B < V_{IG(min)}$$

Typically, V_B is then put between 0·80 and 0·95V_{IG}.

(*b*) Determine R_A so that $I_{K(peak)}$ is not exceeded when a tube is struck with C_K discharged.

$$R_A \geqslant (V_{B(max)} - V_{M(min)})/I_{K(peak)\,(max)} \qquad (4.22)$$

(*c*) Determine R_K to provide the required minimum output voltage, $V_{K(min)}$.

$$R_K \geqslant \frac{R_{A(max)} \,.\, V_{K(min)}}{V_{B(min)} - V_{M(max)} - V_{K(min)}} \qquad (4.23)$$

The maximum value of V_K which may be attained is limited by the necessity of keeping I_K in the region of normal glow. Thus R_K cannot be made much greater than R_A and is often about equal to R_A.

(*d*) Knowing R_K and R_A, use Fig. 4.14 to determine αT.

(*e*) Insert this value of αT in Equation (4.19a) to determine the minimum safe value of C_K.

$$C_{K(min)} = \frac{T_D}{\alpha T \,.\, R_K} \qquad (4.19a)$$

Pulsed-anode Extinguishing

Fig. 4.15 indicates a further method whereby a trigger tube may be extinguished. A negative-going pulse is applied through capacitor C to hold the tube anode below V_M long enough for deionization to occur. The trigger discharge is extinguished by the diode D_1 as described earlier.

If the pulse comprises a negative-going step function of amplitude V_F, then the anode potential of V_1 first falls by V_F and then rises exponentially towards V_B. Thus:

$$V_A = V_B - (V_B - V_M + V_F) \,.\, e^{-t/CR_A}$$

The minimum value of C providing reliable extinguishing is that for which V_A rises to V_M in time t equal to the deionization time, T_D. Thus:

$$(V_B - V_M) = (V_B - V_M + V_F) \,.\, e^{-T_D/C_{(min)}R_A}$$

whence
$$C_{(min)} = \frac{T_D}{R_A} \div \log_e\left(1 + \frac{V_F}{V_B - V_M}\right) \qquad (4.24)$$

In the arrangement shown in Fig. 4.15 (*a*) there is a danger that the quenching pulse may become sufficiently differentiated by C and R_A to cause anode–cathode breakdown on the positive spike produced on the return edge of the pulse. This danger arises only if V_F exceeds $(V_{IG} - V_B)$. It may thus always be avoided by keeping V_F small and $C \,.\, R_A$ large. Alternatively, a catching diode D_2 can be used as indicated to limit the positive swing to a safe value.

This problem does not arise with the arrangement shown in Fig. 4.15 (*b*). Two tubes, V_1 and V_2, are used in complementary manner so that ignition of one applies an extinguishing pulse through C to the anode of the other. When similar tubes are used,

$$V_F = V_B - V_M$$

Thus, from Equation (4.24),

$$C_{(min)} = 1{\cdot}44T_D/R_A \qquad (4.24a)$$

As T_D is normally quoted with reference to an applied anode voltage in excess of V_M, this value of $C_{(min)}$ should be conservative. On the

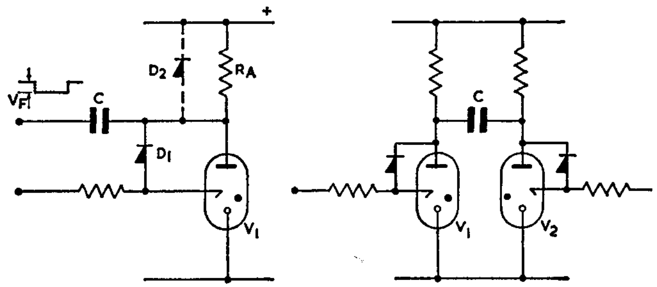


FIG. 4.15. Pulsed-anode extinguishing (*a*) by independent pulse source, and (*b*) by complementary tube, V_2.

other hand, it is based on the assumption that the value of V_M is the same for both tubes. A spread in the values of V_M calls for a larger value of C. This may be calculated rigorously from Equation (4.24), but in practice a sufficiently accurate and conservative estimate of $C_{(min)}$ is given by Equation (4.24b).

$$C_{(min)} = 1{\cdot}5\left(\frac{V_B - V_{M(min)}}{V_B - V_{M(max)}}\right) \cdot \frac{T_D}{R_A} \qquad (4.24b)$$

Extinguishing by Pulsing of H.T.

Any number of tubes may be extinguished simultaneously by pulsing the h.t. supply rail below V_M for a time exceeding the tube deionization time.

For circuits which do not have to operate at high speeds, unsmoothed rectified a.c. of mains frequency may be used for the tube anode supply. If the trigger potential is maintained above V_T the tube will trigger as each cycle of the power supply carries the anode voltage above V_M. The tube will thus carry pulsating current so long as the trigger is held

positive. Provided the supply frequency does not exceed a few hundred cycles per second, the tube will deionize each time the anode supply falls below V_M. Hence the tube fails to restrike on subsequent cycles if the positive trigger bias is removed.

Brierley [11] has described in detail a range of automation circuits based on a refined use of this technique. In this system, ('ACCESS'), the trigger inputs are also in the form of alternate half-sines. A tube will fire only when it receives a trigger input in phase with the alternate half-sines applied to its anode. It is thus possible to pass half-sines of complementary phases through a single wire to control independently two trigger tubes having anode supplies derived from opposite half-cycles of the mains supply. By simple means, the 'ACCESS' system therefore combines the following advantages:

(1) Automatic extinguishing of anode discharge on removal of input.
(2) 'A' and 'B' signals may be transmitted over the same wires.
(3) Tube responds only to 'A' or to 'B', depending on phase of anode supply.
(4) Tube responds to 'A' *and* to 'B' if full-wave supply is provided to anode.
(5) Circuits and power supplies are simple and reliable.

Young [14] has reported further refinements of the system, notably a single-tube circuit for translating a signal from phase 'A' to phase 'B' and a 'NOT' gate. Unfortunately, his circuits depend for their action on the passage of reverse trigger current, and this impairs tube stability.

Circuits operating at higher speeds generally call for h.t. pulses of substantially rectangular waveform. Since the 'off' period must be sufficiently long to ensure deionization, the maximum repetition rate is limited by the deionization time of the tubes employed. As the ionization time of a tube is generally much shorter than its deionization time, the on–off ratio of the h.t. pulse may be quite small. Accordingly, in complicated high-speed circuits several pulse phases may be used, each controlling a separate group of tubes. Pulse generators used for this purpose may be of either the series or the shunt stabilizer type. In a circuit described by Flood and Warman [12] the h.t. supply is through a cathode follower, both grid and cathode of which are dragged to a low voltage when positive-going pulses are applied to the grid of a shunt valve.

Self-Quenching

Just as a cold cathode diode discharge may be made self-quenching by using an appropriate series resistance and shunt capacitance, so either the anode or the trigger discharge of a trigger tube may be made self-quenching by similar means. When the tube strikes, the voltage across the gap falls to its maintaining voltage. The capacitor discharges rapidly, producing heavy ionization of the tube. Consequently, although no further ionization takes place once the gap voltage has fallen to the maintaining voltage, the removal of ions from the gap constitutes a continued flow of current. If this exceeds the current flowing in the series resistor the capacitor will be discharged below the normal maintaining voltage. As the ion current decreases, the capacitor will begin to recharge. Provided that it takes longer than the deionization time T_D to reach the gap voltage for which T_D is quoted, the discharge will not immediately restrike.

Fig. 4.16, due to Crowther and Gimson [13], shows the relation

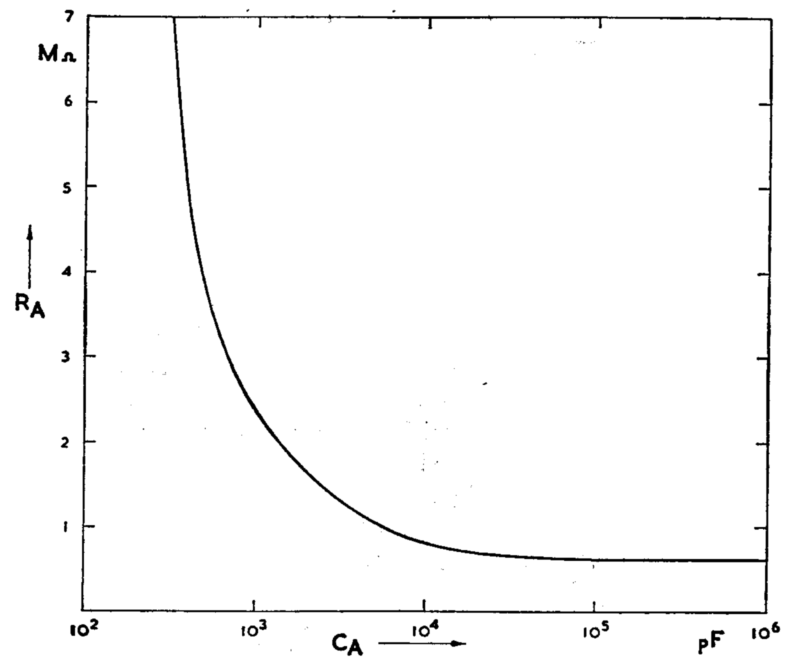


Fig. 4.16. Relation between minimum values of C_A and R_A providing self-quenching of anode–cathode discharge in CV2434.

between C_A and the minimum value of R_A providing self-quenching in the anode gap of a CV2434 (Z803U). When C_A exceeds about 0·01 μF. the minimum value of R_A is substantially constant and is determined by the ion current mentioned above. For smaller values of C_A it is necessary to increase R_A to ensure the gap voltage does not recover in less than the time T_D.

With large values of C_A (or C_T in the trigger circuit) a resistor must be connected in the capacitor discharge path to limit the peak discharge

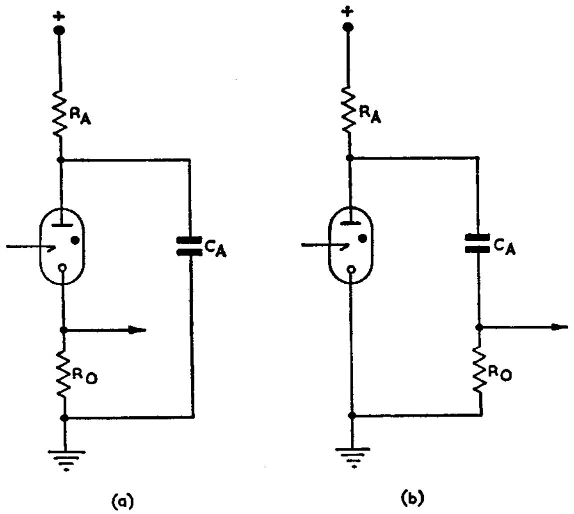


Fig. 4.17. Tube operating in self-quenching mode to generate low-impedance pulses of (*a*) positive and (*b*) negative polarity.

current to a value which will not damage the tube. The limiting value of trigger current is much lower than the limiting value of anode current. Accordingly, limiting resistors are needed with values of C_T much smaller than the values of C_A which may be used without series resistors.

A series resistor is also used to develop an output pulse at low impedance. The pulse polarity may be positive or negative, depending on whether the resistor is connected in the position shown in Fig. 4.17 (*a*) or in that of Fig. 4.17 (*b*). The pulse amplitude cannot exceed ($V_{IG} - V_M$), and in practice is significantly less. During the pulse the output impedance is low, being given approximately by the parallel impedance of R_0 and the tube. The tube impedance is indeterminate, but may be taken to be somewhat less than 1 kΩ. Thus a self-quenching discharge may be used to generate sharp, low-impedance pulses from which cathode

triggering may be effected according to Fig. 4.10 (*b*). Alternatively, a self-quenching discharge can be used to apply pulses simultaneously to the triggers of a number of tubes, as in pulse-plus-bias triggering.

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CHAPTER FIVE

Trigger Tube Circuits and Applications

Some basic circuit techniques were discussed in the preceding chapter. Here it will be shown how they are combined with further techniques to solve specific problems.

Relay Circuits

Trigger tubes are widely used to indicate change of state of an input quantity. Typical examples are various types of alarm and protection circuits responding to touching of a contact [1], failure of a flame [2, 3], or breaking of a thread [3]. All these devices operate from a relatively large change of input level. Circuits differ considerably, depending whether or not they are to reset automatically when the input returns to its normal state. In some cases the need for a fail-safe design demands that the tube be normally conducting. A cessation of current flow then indicates a fault condition.

Touch-circuit Design

The trigger circuit shown in Fig. 5.1 is typical of that used in touch-operated devices. Contact X may be a probe supported over the surface

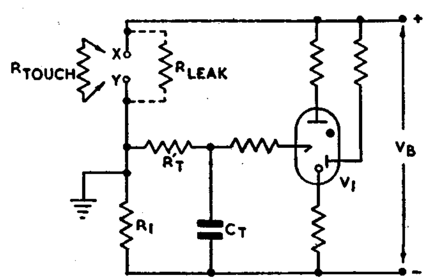

Fig. 5.1. Touch-operated relay, suitable for liquid level control.

of a liquid. When the level rises so that the surface touches X, a circuit, R_{touch}, is formed through the liquid to a submerged electrode, Y. In another application X is the wire of an electric fence. When an animal

forms a circuit between fence and ground the trigger tube fires and applies a high-voltage pulse to the wire.

In designing circuits of this type, R_1 must be chosen to satisfy two requirements:

(1) In the absence of R_{touch} the minimum leakage resistance, R_{leak}, across X and Y must not fire V_1 with the maximum supply voltage V_B and the minimum trigger breakdown voltage, V_T.

(2) With the highest value of R_{touch} and no leakage, firing must occur with V_B at a minimum and V_T at a maximum.

These two conditions give respectively:

$$V_{B(max)} \cdot \frac{R_1}{R_1 + R_{leak}} < V_{T(min)} \tag{5.1}$$

and

$$V_{B(min)} \cdot \frac{R_1}{R_1 + R_{touch(max)}} > V_{T(max)} \tag{5.2}$$

Combining these two relations, upper and lower limits for R_1 are given by Relation (5.3).

$$R_{leak} \cdot \frac{V_{T(min)}}{V_{B(max)} - V_{T(min)}} > R_1 > R_{touch(max)} \cdot \frac{V_{T(max)}}{V_{B(min)} - V_{T(max)}} \tag{5.3}$$

Assuming Relation (5.3) can be satisfied, R_1 should be chosen to provide approximately the same factor of safety in each half of the relation. For example, if the upper and lower limits for R_1 differ by a factor of 10, R_1 should be made about one-third of the upper limiting value, not one-half. Thus

$$R_{1(optimum)} = (R_{1(max)} \times R_{1(min)})^{\frac{1}{2}} \tag{5.4}$$

R_T' is required to limit the trigger current in the event of a short circuit between X and Y. Thus

$$R_T' > \frac{V_{B(max)} - V_N}{I_{T(max)}} \tag{5.5}$$

A check should be made to ensure that Relation (4.7) is satisfied, remembering that R_T' is increased by the resistance of R_1 and R_{touch} in parallel. Hence from Relation (4.7),

$$R_T' + \frac{R_1 \cdot R_{touch(max)}}{R_1 + R_{touch(max)}} \ll \frac{V_T}{I_P} \tag{5.6}$$

The same design procedure may be applied also to photoelectric control circuits in which a photoconductive cell is connected between X and Y.

EXAMPLE 5.1 Touch Circuit Relay

A touch-operated intruder alarm is required to respond to contact to earth through 100 kΩ. Leakage resistance may be as low as 1 MΩ. Using a CV2434 (Z803U), determine R_1 and R_T' for $V_B = 240\text{ V} \pm 20\text{ V}$, $V_T = 128$ to 137 V, $I_P = 3 \times 10^{-8}$ A. Restrict trigger current to 0·5 mA. $V_N \approx 90$ V.

(*a*) From Relation (5.3), evaluate the upper limit of R_1.

$$R_{1(\max)} = R_{\text{leak}} \cdot \frac{V_{T(\min)}}{V_{B(\max)} - V_{T(\min)}} \tag{5.3a}$$

$$= 10^6 \cdot \frac{128}{220 - 128} = \mathbf{1{\cdot}39\ M\Omega}$$

(*b*) From Relation (5.3), evaluate the minimum value of R_1.

$$R_{1(\min)} = R_{\text{touch}} \cdot \frac{V_{T(\max)}}{V_{B(\min)} - V_{T(\max)}} \tag{5.3b}$$

$$= 10^5 \cdot \frac{137}{260 - 137} = \mathbf{111\ k\Omega}$$

(*c*) Choose optimum value of R_1.

$$R_{1(\text{opt})} = (R_{1(\max)} \times R_{1(\min)})^{\frac{1}{2}} \tag{5.4}$$

$$= (1{\cdot}39 \times 10^6 \times 111 \times 10^3)^{\frac{1}{2}}$$

$$= 392\text{ k}\Omega\text{, say } \mathbf{390\ k\Omega}$$

(*d*) Determine R_T' to limit the maximum trigger current.

$$R_T' > \frac{V_{B(\max)} - V_N}{I_{T(\max)}} \tag{5.5}$$

$$= \frac{260 - 90}{0{\cdot}5} = 300\text{ k}\Omega\text{, say } \mathbf{330\ k\Omega}$$

(*e*) Check that the trigger circuit resistance is not excessive.

$$R_T' + \frac{R_1 \cdot R_{\text{touch}}}{R_1 + R_{\text{touch}}} = 330 + \frac{390 \cdot 100}{390 + 100} = 408\text{ k}\Omega$$

$$\frac{V_T}{I_P} = \frac{132}{3 \times 10^{-8}} = \mathbf{4{,}400\ M\Omega}$$

Hence Relation (5.6) is satisfied handsomely and the design is acceptable.

(*f*) From manufacturer's data, $C_T = 2{,}700$ pF.

Solution: $R_1 = 390$ kΩ, $R_T' = 330$ kΩ, $C_T = 2{,}700$ pF. No close tolerances required.

Degenerative Voltage Stabilization

Using a circuit of the type shown in Fig. 5.2, an amplifier may be constructed in which a trigger tube provides a pseudo-continuous control characteristic. The values of C_1 and R_1 are chosen so that an input E_T exceeding V_T produces relaxation oscillations (p. 38) in the trigger circuit. In the anode circuit C_2 and R_2 are chosen so that when the trigger circuit discharge occurs, an anode–cathode discharge is produced which is also self-quenching. By making C_2R_2 small enough to allow full

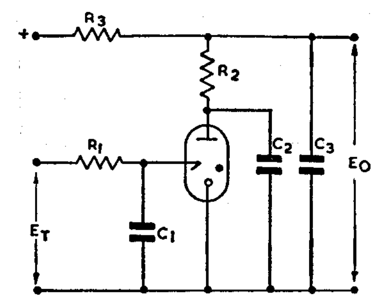

Fig. 5.2. Amplifier providing pseudo-continuous control.

recharging between successive discharges, a linear relation is obtained between mean anode current and the frequency of firing of the trigger circuit. C_3R_3 is made sufficiently large to smooth the current pulsations through R_3 and an output voltage, E_0, may be obtained across C_3 which is a function of the input voltage E_T.

Goulding [4] has shown that, using a suitable tube, an effective 'mutual conductance' of 500 μA/V should be obtainable. In practice, he finds statistical variations in the trigger potential reduce this to about 150 μA/V. This is nevertheless sufficient to make practicable a shunt stabilizer in which a trigger tube is used as an amplifier.

Goulding describes two circuits similar to that of Fig. 5.3. A potentiometer formed by R_X and R_Y applies part of the output E_0 to the trigger. The trigger striking potential, V_T, serves as the voltage reference. If the output rises above $(1 + R_Y/R_X) \cdot V_T$, trigger circuit oscillations occur and the consequent flow of anode current produces a voltage drop across R_3 restricting the rise in output.

A full analysis of this circuit would appear to be difficult, but Goulding describes alternative designs providing (*a*) 360–420 V at 0–3 μA from a 600–800-V d.c. supply, and (*b*) 151–165 V at 0–80 μA with an output

impedance of 10 kΩ. For battery-operated Geiger counters, this type of circuit proved attractive in its economy of current consumption and in having an output voltage adjustable without adverse effect on output impedance.

A circuit arrangement due to Kerr and van Vlodrop [5] bears a

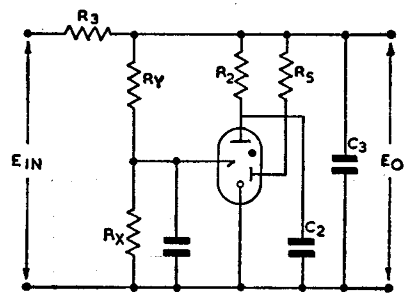

Fig. 5.3. Degenerative voltage stabilizer, suitable for low currents at comparatively high voltages.

superficial resemblance to that of Goulding, but is superior as regards both 'designability' and performance. It allows some tens of milliamperes to be delivered at an output impedance considerably lower than that provided by a conventional stabilizing diode.

Consider first the performance of the basic circuit shown in Fig. 5.4

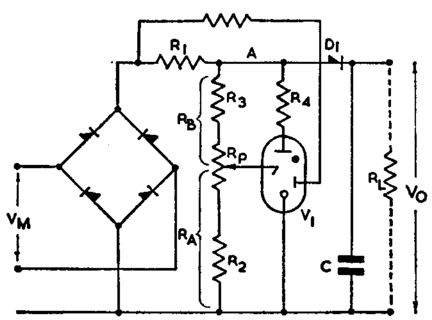

Fig. 5.4. Degenerative trigger tube shunt stabilizer providing current $\approx I_K$ and voltage $\leqslant V_{IG}$.

if the trigger tube V_1 is omitted. The bridge rectifier supplies the load R_L through the series resistance R_1 and diode D_1. Assuming $(R_A + R_B) \gg R_1$, the waveforms obtained across the rectifier bridge and the output would be as shown at (*a*) and (*b*) respectively in Fig. 5.5. The rectifier bridge provides an output comprising a train of half-sines of amplitude $V_m\sqrt{2}$.

Between time t_0 and t_1 on each half-sine, the bridge output is less than the voltage remaining across C. Accordingly, D_1 is non-conducting during this time. At t_1, D_1 conducts and C recharges through R_1. Current flows through R_1 over only a fraction of the complete cycle. While it flows, therefore, the current substantially exceeds the mean output current. Although R_1 is not large, it will give rise to an appreciable voltage drop and immediately after t_1 the output voltage curve (*b*) rises less rapidly than curve (*a*). Between t_1 and t_2 the output capacitor is

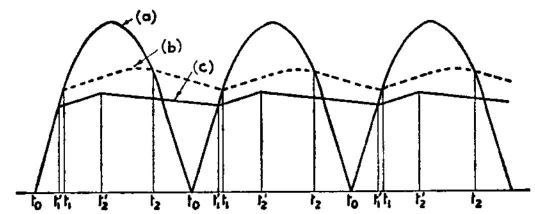

FIG. 5.5. Voltage waveforms corresponding to Fig. 5.4.

charging, but at t_2 the bridge output falls below the voltage on C and the diode D_1 cuts off once more. The output voltage then falls as C discharges through R_L, and this continues until time t_1 is reached on the following cycle.

If, now, the circuit is considered with the trigger tube V_1 in position the broken curve (*b*) is replaced by the full line (*c*). Because the trigger tube stabilizes the output voltage at a reduced value, the time t_1 at which D_1 conducts is advanced to t_1'. Thereafter C charges until time t_2' at which the potential divider formed by R_A and R_B applies a trigger voltage sufficient to fire V_1. The anode current drawn by V_1 then holds the potential at point A below that of the capacitor C. Thus D_1 is cut off and the output voltage falls slowly as C discharges through R_L until on the next cycle D_1 conducts again.

At the end of each cycle the trigger tube discharge extinguishes as the output voltage of the rectifier bridge falls below V_M, the anode maintaining voltage. Thus V_1 does not conduct again until time t_2' is reached on the following cycle.

It is evident that, provided the circuit has been designed successfully to work according to the scheme described, the peak value of output voltage (neglecting the drop across D_1 when conducting) is given by

$$V_{0(\text{peak})} = \frac{R_A + R_B}{R_A} \cdot V_T \qquad (5.7)$$

Within certain limits, therefore, $V_{0(peak)}$ is independent of V_M, but may be varied by adjusting R_A/R_B – by inserting a potentiometer R_P between R_2 and R_3 as indicated in Fig. 5.4, for example. Also it will be seen from Equation (5.7) that $V_{0(peak)}$ is independent of the load current (again within certain limits to be noted later). The ripple in the output voltage depends on the value of capacitance it is practicable to use for C. The mean output voltage falls as the ripple increases, and the regulation of the stabilizer therefore depends primarily on the value of C.

When C is made large the amplitude V_r of the ripple is small and the discharge of C may therefore be considered approximately linear. Consequently, the mean output voltage may be written as

$$V_{0(mean)} = V_{0(peak)} - \tfrac{1}{2}V_r \tag{5.8}$$

But

$$V_r = I_L \, . \, t_d/C \tag{5.9}$$

where t_d = discharge time of C.

The output impedance of the stabilizer is given by

$$R_0 = -\frac{dV_{0(mean)}}{dI_L}$$

Hence, from Equations (5.8) and (5.9),

$$R_0 \approx \tfrac{1}{2}t_d/C \tag{5.10}$$

(This ignores the variation of t_d with I_L, but leads to a slightly pessimistic result.)

Putting $t_d \approx 1/4f$, where f is the supply frequency in cycles per second,

$$R_0 \approx \frac{1}{8fC} \tag{5.10a}$$

Hence, if $C = 50$ μF and $f = 50$ c/s, an output impedance of about 50 Ω may be expected. In practice, the variation of t_d with I_L lowers this figure and, with the values considered above, $R_0 \approx 30$ Ω. The corresponding amplitude of the ripple in the output is 4 V at 10 mA output current.

Determination of component values can be a complex problem if tolerances of V_m and V_0 are to be accommodated without making V_m needlessly high and without exceeding the limiting current ratings of V_1. Although it makes certain approximations, the following treatment leads to a design procedure which is quite straightforward and sufficiently accurate for most purposes.

First it is assumed that C is made so large that the ripple voltage becomes negligible compared to V_0 and V_m. Then, for given values of

V_m, V_0, and R_1, the maximum output current at which V_0 can be maintained is given by

$$I_{0(max)} = \frac{1}{\pi}\left\{\int_a^b \frac{V_m\sqrt{2}}{R_1} \, . \sin\theta \, . \, d\theta - \int_a^b \frac{V_0}{R_1} \, . \, d\theta\right\} \tag{5.11}$$

where a and b are the values of θ, as indicated in Fig. 5.6, at which the instantaneous value of V_m is equal to V_0,

i.e.

$$a = \sin^{-1}\frac{V_0}{V_m\sqrt{2}} \tag{5.12}$$

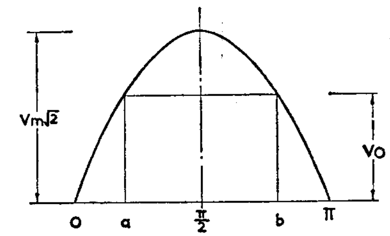

Fig. 5.6. Inter-relation of rectifier conduction angle (a to b), supply voltage (V_m) and maximum output voltage (V_0) across reservoir capacitor.

From Equation (5.11),

$$I_{0(max)} = \frac{2}{\pi R_1}\left\{V_m\sqrt{2}\int_a^{\pi/2} \sin\theta \, . \, d\theta - \int_a^{\pi/2} V_0 \, . \, d\theta\right\}$$

whence

$$I_{0(max)} = \frac{2}{\pi R_1}\left\{(2V_m^2 - V_0^2)^{\frac{1}{2}} - V_0\left(\frac{\pi}{2} - \sin^{-1}\frac{V_0}{V_m\sqrt{2}}\right)\right\} \tag{5.13}$$

Now, since this is a shunt stabilizer, removal of a load requires that $I_{0(max)}$ shall flow through the shunt tube, V_1. Hence $I_{0(max)}$ must not exceed the maximum average cathode current of V_1. The dangerous condition is, of course, when the a.c. supply has risen to $(1 + r_1) \, . \, V_m$ and when a low value of V_0 is required. (The values of R_1 and R_4 must be chosen to allow V_1 to hold the point A below the minimum output voltage. In the absence of an external load, the current drawn by V_1 is thereafter substantially independent of the set value of V_0.) In the no-load condition,

$$I_{K(av)\,(max)} \geqslant \frac{2}{\pi R_1}\left\{([(1 + r_1)V_m\sqrt{2}]^2 - V_{0(min)}^2)^{\frac{1}{2}} - V_{0(min)} \, . \left(\frac{\pi}{2} - \sin^{-1}\frac{V_{0(min)}}{(1 + r_1)V_m\sqrt{2}}\right)\right\} \tag{5.13a}$$

Insertion of known values of r_1, V_m, V_0, and $I_{K(av)\,(max)}$ leads to evaluation of R_1. If V_m does not greatly exceed V_0, however, the peak cathode current, $I_{K(pk)}$, might exceed the maximum permissible value, $I_{K(pk)\,(max)}$.

If $V_m = k \,.\, V_0$, the curve of p versus k in Fig. 5.7 simplifies a check that the limit is not exceeded, i.e. that

$$I_{K(pk)} = p \,.\, I_{K(av)} \leqslant I_{K(pk)\,(max)} \tag{5.14}$$

With most trigger tubes there is little danger of exceeding this restric-

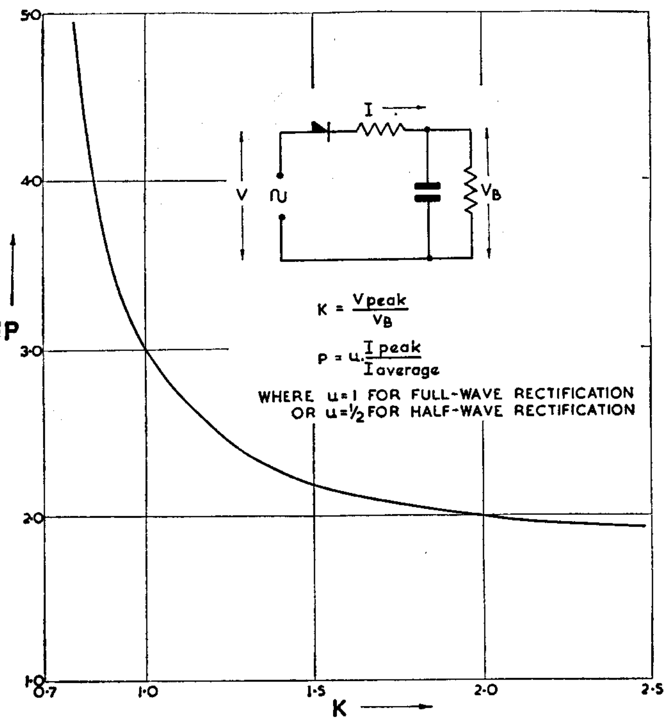

Fig. 5.7. Design curve for circuits operating on rectified sinusoidal a.c.

tion provided the tube is operating on every half-cycle of the a.c. supply. If the rectifier bridge is replaced by a half-wave rectifier, however, the value of $I_{K(av)\,(max)}$, *averaged over the conducting half-cycle only*, is doubled. The value of $I_{K(av)\,(max)}$ is not doubled, however, and the acceptable value of p is thus halved. As a result, the value of k may have to be raised in order to reduce p.

When the value of R_1 has been determined for given values of V_m

and V_0 it is necessary to know that the requisite value of $I_{0(max)}$ can be supplied under the most adverse conditions, i.e. $(1 - r_2) \,.\, V_m$ and $V_{0(max)}$.

If one assumes, first, that $V_{0(max)} = V_{0(min)} = V_0$, then

$$I_{pk(Vm\ low)} = \frac{(1 - r_2)V_m\sqrt{2} - V_0}{R_1} \tag{5.15}$$

and

$$I_{pk(Vm\ high)} = \frac{(1 + r_1)V_m\sqrt{2} - V_0}{R_1} \tag{5.16}$$

Combining these expressions,

$$I_{av(Vm\ low)} = \frac{1}{p} \,.\, I_{pk(Vm\ low)} = \frac{1}{p} \,.\, \frac{(1 - r_2) \,.\, V_m\sqrt{2} - V_0}{(1 + r_1) \,.\, V_m\sqrt{2} - V_0} \,.\, I_{pk(Vm\ high)}$$

$$\therefore\ I_{av(V_m\ low)} = q \,.\, I_{av(V_m\ high)} \approx \frac{(1 - r_2) \,.\, k\sqrt{2} - 1}{(1 + r_1) \,.\, k\sqrt{2} - 1} \,.\, I_{av(V_m\ high)} \tag{5.17}$$

In designing a stabilizer for a constant output voltage V_0 and maximum current I_0, q is first determined as $I_{0(max)}/I_{K(av)\,(max)}$, where $I_{K(av)\,(max)}$ relates to the trigger tube to be used. The lowest value of k is selected which will conform to the value of q and the supply voltage tolerance $\pm r$. The value of p corresponding to k is determined from Fig. 5.7 and R_1 is calculated from Equation (5.18).

$$R_1 = \frac{(1 + r_1) \,.\, V_m\sqrt{2} - V_0}{p \,.\, I_{K(av)\,(max)}} \tag{5.18}$$

When V_0 is to be adjustable from $V_{0(min)}$ to $m \,.\, V_{0(min)}$ a higher value for k is required. The most difficult condition now corresponds to an input of $(1 - r_2) \,.\, V_m$ and an output of $m \,.\, V_{0(min)}$. The maximum average output current in this condition is given by

$$I_{0(av)\,(max)} \approx \frac{1}{p} \,.\, I_{pk}$$

$$= \frac{1}{p} \,.\, \frac{(1 - r_2) \,.\, V_m\sqrt{2} - \mathrm{m} \,.\, V_0}{R_1} \tag{5.19}$$

Combining Equations (5.16) and (5.19),

$$\frac{I_{0(av)\,(max)}}{I_{pk(V_m\ high)}} = \frac{1}{p} \,.\, \frac{(1 - r_2) \,.\, V_m\sqrt{2} - \mathrm{m} \,.\, V_0}{(1 + r_1) \,.\, V_m\sqrt{2} - V_0}$$

The minimum value of V_m corresponds to the limiting condition of Relation (5.14).

$$\therefore\ q' \equiv \frac{I_{O(av)\ (max)}}{I_{K(av)\ (max)}} = \frac{(1 - r_2) \,.\, V_m\sqrt{2} - m \,.\, V_O}{(1 + r_1) \,.\, V_m\sqrt{2} - V_O}$$

$$\simeq \frac{(1 - r) \,.\, k'\sqrt{2} - m}{(1 + r) \,.\, k'\sqrt{2} - 1} \tag{5.20}$$

where $\qquad r = \frac{1}{2}(r_1 + r_2)$

and $\qquad k' = \frac{V_m'}{V_{O(min)}} = \left(1 + \frac{r_1 - r_2}{2}\right)\frac{V_m}{V_{O(min)}} \qquad (5.21)$

From Equation (5.20),

$$k' = \frac{m - q'}{[(1 - r) - q'(1 + r)]} \cdot \frac{1}{\sqrt{2}} \tag{5.20a}$$

Once k' has been determined, V_m, p, and R_1 may be found as for the case in which V_O is constant. R_4 may be determined from

$$R_4 = \frac{V_{O(min)} - V_M}{p \,.\, I_{K(av)\ (max)}} \tag{5.22}$$

Finally,

$$R_P = m \,.\, R_2 \tag{5.23}$$

$$R_3 = \left(\frac{V_{O(min)}}{V_T} - 1\right) . \, mR_2 \tag{5.24}$$

and so that the potentiometer chain shall not appreciably load R_1, R_A must be made as large as possible. It is not practicable to use a trigger–cathode capacitor, however, and hence the trigger circuit impedance is restricted by the limitations imposed by Relations (4.4) and (4.5). Thus

$$R_T = \frac{R_A R_B}{R_A + R_B} < \frac{V_T - V_N}{I_T} \tag{5.25}$$

When $V_O = V_T$, $R_B = R_A$. The Relation (5.25) simplifies to

$$R_{A(max)} = 2\left(\frac{V_T - V_N}{I_T}\right) \tag{5.25a}$$

Where I_T = transfer current for $V_A = 2V_T$.

It will be seen also that, for lower values of V_O, R_B decreases and I_T increases.

Consequently, Relation (5.25a) is approximately true for all practical cases.

If $R_{A(max)}$ is exceeded the tube will not trigger at the required anode voltage. If R_A approaches within an order of $R_{A(max)}$ triggering becomes erratic due to hysteresis effects (p. 67). A practical value for R_A is thus given by

$$R_A \simeq \frac{V_T - V_N}{10 \,.\, I_T} \tag{5.25b}$$

where I_T = transfer current for $V_A = 2V_T$.

No mention has yet been made of resistor tolerances. When these are considered it is found that a tolerance $\pm w$ demands that R_P be increased by the factor $(1 + w)/(1 - w)$ while R_3 and q' are reduced by the same factor.

Performance Limitations

The foregoing treatment should not be taken to indicate that, by using sufficiently large values of C, any desired degree of stabilization is possible. When C is increased to reduce the ripple to a very small value the trigger potential rises very slowly once the diode conducts. It can then happen that, when the trigger tube is lightly loaded (i.e. the stabilizer is operating near to full load), the trigger tube will conduct only on alternate half-cycles of the supply. The ripple frequency will then be that of the supply (instead of double the supply frequency) and the ripple voltage will be approximately twice the calculated value.

'Trigger hysteresis' (p. 67) can lead to significant regulation effects. If, for example, the supply voltage is abruptly increased the immediate effect may be a slight rise in V_O. At the new supply voltage, however, the trigger tube is required to carry an increased current. The consequently increased heating of the tube produces a depression of V_T, and hence of V_O. Thus trigger-tube shunt stabilizers may prove to be slightly over-compensated in respect of supply variations. On the other hand, the same mechanism leads to an observed output impedance lower than the calculated value – provided observations are delayed for a time equal to the thermal time-constant of the tube (usually less than 10 seconds).

Limitations such as these preclude the use of the trigger tube shunt stabilizer where the very highest stability is required. On the other hand, a stability of $\pm 2\%$ is well within its capabilities.

Voltage and Current Limitations

The simple stabilizer of Fig. 5.4 is restricted as to the voltage and current ranges it can cover. V_O cannot be less than V_T nor more than the anode breakdown voltage, V_{IG}. The output current is restricted to less than $I_{K(av)\ (max)}$. When necessary, any of these limitations may be removed by modified circuit arrangements.

Low values of V_O may be obtained by using two circuits similar to Fig. 5.4. The output is taken as the difference between their positive outputs. The circuit providing the lower output voltage must be provided with a resistive load drawing a current at least equal to the load current drawn between the positive outputs of the two circuits.

For high voltages two or more stabilizers may be superimposed. The

circuit of Fig. 5.8 (due to Kerr and van Vlodrop [5]) effects some economy by using a half-wave supply to either trigger tube, the diodes D_1 and D_2 protecting the anodes of V_1 and V_2 against the application of reverse voltage. Since either tube conducts only on alternate half-cycles of the supply, the maximum value of $I_{K(av)}$ over this half-cycle is $2I_{K(av)\,(max)}$. As a result, the permissible maximum value of p is only $\frac{1}{2}I_{K(pk)\,(max)}/I_{K(av)\,(max)}$ and a rather high value of V_m becomes necessary if the load current is to approach $I_{K(av)\,(max)}$. In many high-voltage

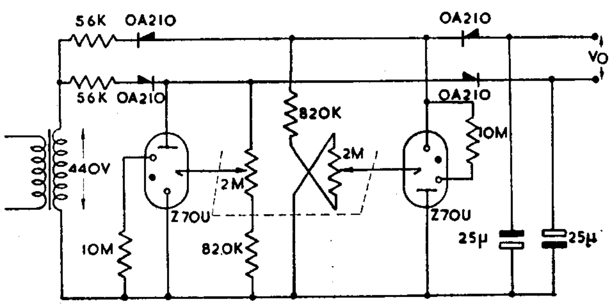

Fig. 5.8. Stabilized supply using voltage doubling.

applications the load current is so small that this is not a serious limitation.

Consideration of Fig. 5.8 will show that across either V_1 or V_2 a waveform of constant peak–peak amplitude is obtained. Consequently, still higher voltages may be obtained (at still lower currents) by using such a waveform to drive a voltage-multiplying rectifier. An example due to Kerr and van Vlodrop is shown in Fig. 5.9. Here a voltage multiplier is used to increase further the output, which has already been doubled by using two tubes head to tail. Design procedure for such a circuit follows that given earlier, with the following provisos:

(1) If voltage multiplication is used, the maximum value of I_0 must be less than $\frac{2}{s} \cdot I_{K(av)\,(max)}$, where s is the overall voltage multiplication factor.

(2) The output impedance is increased by a factor of about $2s$.

(3) Since two tubes are used, conducting on alternate half-cycles, determination of R_1 proceeds as for the full-wave case previously described, but with a doubling of the true value of $I_{K(av)\,(max)}$ wherever it occurs in the calculation.

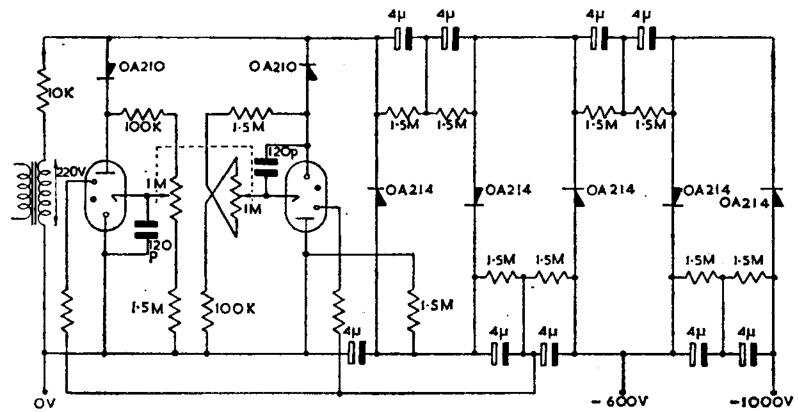

Fig. 5.9. High-voltage supply providing adjustable stabilized output.

Fig. 5.10 shows an arrangement permitting the use of a number of trigger tubes in parallel so that quite large currents may be controlled. The potentiometer supply to each trigger electrode is returned to the cathode of an adjacent tube. Thus when any one tube fires, the rise in cathode potential triggers the next tube. As the interconnexion forms a

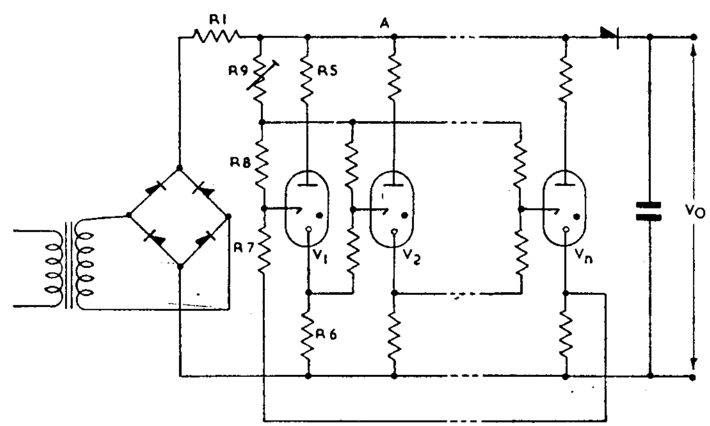

Fig. 5.10. Stabilized supply for high output currents.

closed ring, all the tubes will be triggered almost simultaneously. Equality of current sharing requires that the tolerance of the anode-maintaining voltage, V_M, shall be small compared with $(V_0 - V_M)$. Accordingly, the circuit is best suited to applications where V_0 is not much less than V_{IG}.

In order to pass a large pulse from cathode to adjacent trigger, it is necessary that $R_7 \gg R_8$. This implies a relatively large value of V_0.

Design of the parallel-tube circuit using n tubes is based on that of a single-tube circuit where $I_{0(max)}$ has been reduced by the factor n. The value of R_1 so obtained must then be divided by n. R_4 is split into two parts, R_5 and R_6, so that a pulse shall be passed on to the following tube as large as consistent with avoiding the passage of reverse trigger current. The most exacting case is that in which firing of a single tube is insufficient to bring the potential of point A below V_0. In this case,

$$V_K = (V_0 - V_M) \, . \, R_6/(R_5 + R_6) \tag{5.26}$$

The maximum value of V_K which can be used without causing reverse trigger current to flow is then given by

$$V_{K(max)} = V_T - V_N \tag{5.27}$$

Hence the maximum value of R_6 is given by

$$R_{6(max)} = \frac{V_T - V_N}{(V_{0(min)}/s) - V_M} \cdot R_4 \tag{5.28}$$

Provided $R_7 \gg R_6$,

$$R_8 = \left(\frac{V_{0(min)}}{V_T} - 1\right) . \, R_7 \tag{5.29}$$

and

$$R_9 = \frac{1}{n}\left\{\left(\frac{V_{0(max)}}{V_T} - 1\right) . \, R_7 - R_8\right\} \tag{5.30}$$

Allowing for component tolerances, these expressions become

$$R_8 = \left(\frac{V_{0(min)}}{V_{T(max)}} - 1\right) . \, R_7 \cdot \frac{(1 - w)}{(1 + w)} \tag{5.29a}$$

and
$$R_9 = \frac{1}{n} \cdot \left\{\left(\frac{V_{0(max)}}{V_{T(min)}} - 1\right) . \, R_7 - R_8\right\} \cdot \frac{(1 + w)}{(1 - w)} \tag{5.30a}$$

Design Procedure for Trigger Tube Shunt Stabilizer

(*a*) Set out the following design data:
- Maximum nominal output voltage, $V_{0(max)}$
- Minimum nominal output voltage, $V_{0(min)}$
- Maximum average output current (averaged over one cycle of a.c. supply), $I_{0(av)\,(max)}$
- Fractional tolerances, $+r_1$ and $-r_2$, on a.c. supply voltage

- Fractional tolerance, $\pm w$, on resistors
- Maximum acceptable output impedance, R_0
- Maximum acceptable output ripple voltage, V_r

(*b*) Tentatively select a trigger tube on which to base design and set out the following data relating to it:
- Anode breakdown potential, V_{IG}
- Anode maintaining potential, V_M
- Nominal trigger striking potential, V_T
- Maximum trigger striking potential, $V_{T(max)}$
- Minimum trigger striking potential, $V_{T(min)}$
- Trigger maintaining potential, V_N
- Transfer current (at $V_A = 2V_T$), I_T
- Maximum permissible average cathode current, $I_{K(av)\,(max)}$
- Maximum permissible peak cathode current, $I_{K(pk)\,(max)}$

(*c*) Decide whether voltage multiplication is necessary.

If $V_{0(min)}$ and $V_{0(max)}$ both lie within the limits $V_{T(max)}$ and V_{IG}, multiplication is not necessary, i.e. $s = 1$.

If they are above these limits, choose the voltage multiplication factor, s, as an integer so that $V_{0(min)}$ and $V_{0(max)}$ lie between $s \, . \, V_{T(max)}$ and $s \, . \, V_{IG}$.

If s is an even number, use a voltage-doubling arrangement, as in Fig. 5.8. Follow with further voltage multiplication by a factor $s/2$ as necessary, on the lines indicated in Fig. 5.9.

(*d*) Decide whether tubes are to operate from half-wave rectified supply ($u = \frac{1}{2}$) or from full-wave rectified supply ($u = 1$). (It is preferable to put $u = 1$, but this raises the cost of the power supply, particularly if it is a voltage multiplier.)

(*e*) Evaluate

$$m = \frac{V_{0(max)}}{V_{0(min)}}$$

(*f*) Choose the number of tubes, n, to operate in parallel and evaluate q' so that

$$q' = \frac{1}{n} \cdot \frac{I_{0(av)\,(max)}}{I_{K(av)\,(max)}} < 1{\cdot}0$$

Usually it will be found necessary to choose n so that $q' < 0{\cdot}65$.

(*g*) Evaluate

$$r = \tfrac{1}{2} \, . \, (r_1 + r_2)$$

(*h*) Determine k' from Equation (5.20b),

$$k' = \frac{\mathrm{m}(1-w) - q'(1+w)}{(1-r)(1-w) - q'(1+r)(1+w)} \cdot \frac{1}{\sqrt{2}} \qquad (5.20\text{b})$$

(*j*) From Fig. 5.7, find the value of p corresponding to k'.

(*k*) Check that

$$p < u \cdot \frac{I_{K(pk)\,(max)}}{I_{K(av)\,(max)}}$$

If this is not so, either change from half-wave to full-wave rectification (i.e. from $u = \frac{1}{2}$ to $u = 1$) or choose the lowest value of k' satisfying this relation.

(*l*) Determine the nominal value of V_m.

$$V_m = \frac{k'}{s} \cdot V_{0(min)} \Big/ \left(1 + \frac{r_1 - r_2}{2}\right) \qquad (5.21\text{a})$$

(*m*) From Equation (5.18a), determine R_1,

$$R_1 = \frac{u}{n} \cdot \frac{(1 + r_1) \, . \, V_m\sqrt{2} - V_{0(min)}/s}{p \, . \, I_{K(av)\,(max)}} \qquad (5.18\text{a})$$

(*n*) From Equation (5.22a) determine R_4,

$$R_4 = u \cdot \frac{(V_{0(min)}/s) - V_M}{p \, . \, I_{K(av)\,(max)}} \qquad (5.22\text{a})$$

(*o*) (i) *If* $n = 1$, $R_6 = 0$.
(ii) If $n > 1$, determine R_6 in Fig. 5.10 such that

$$R_6 = \frac{V_T - V_N}{(V_{0(min)}/s) - V_M} \cdot R_4 \qquad (5.28)$$

Then $R_5 = R_4 - R_6$.

(*p*) Choose

$$R_7 \approx \frac{V_T - V_N}{10 \, . \, I_T} \qquad (5.25\text{b})$$

(*q*) Determine R_8 from Equation (5.29a),

$$R_8 = \left(\frac{V_{0(min)}}{s \, . \, V_{T(max)}} - 1\right) \cdot R_7 \cdot \frac{(1-w)}{(1+w)} \qquad (5.29\text{a})$$

(*r*) Determine R_9 from Equation (5.30a),

$$R_9 = \frac{1}{n} \cdot \left\{\left(\frac{V_{0(max)}}{s \, . \, V_{T(min)}} - 1\right) . \, R_7 - R_8\right\} \cdot \frac{(1+w)}{(1-w)} \qquad (5.30\text{a})$$

If necessary, adjust R_7, R_8, and R_9 by equal percentages to allow convenient values to be adopted.

(*s*) From Relation (5.10b), determine C to meet requirements for output impedance, R_0.

$$C \geqslant \frac{s}{8fR_0} \cdot \left(\frac{2-u}{u}\right) \qquad (5.10\text{c})$$

(*t*) From Relation (5.9b), determine C to meet requirements for output ripple voltage, V_r.

$$C \geqslant \frac{1}{4f} \cdot \left(\frac{2-u}{u}\right) \cdot \frac{I_{0(max)}}{V_r} \qquad (5.9\text{b})$$

(*u*) Adopt a value for C not less than the larger value given by Relations (5.10c) and (5.9b) above.

EXAMPLE 5.2 Single-tube Trigger-tube Shunt Stabilizer

Design a trigger-tube shunt stabilizer to operate from a 50-c/s supply at 240 V r.m.s. (+6%, −10%) and to deliver 0–10 mA at any d.c. output voltage between 145 and 290 V. Output impedance to be less than 30Ω, ripple less than 1%. Use 5% tolerance resistors.

(*a*) Setting out design data:

$V_{0(max)} = 290$ V	$I_{0(av)\,(max)} = 10$ mA	$f = 50$ c/s
$V_{0(min)} = 145$ V	$r_1 = 0{\cdot}06$	
$V_{r(max)} = 1{\cdot}45$ V	$r_2 = 0{\cdot}10$	
$R_{0(max)} = 30\ \Omega$	$w = 0{\cdot}05$	

(*b*) Basing design on CV2434,

$V_{IG} = 290$ V	$I_{K(av)\,(max)} = 25$ mA
$V_M = 105$ V	$I_{K(pk)\,(max)} = 100$ mA
$V_T = 132$ V	$I_T = 0{\cdot}045$ mA
$V_{T(max)} = 137$ V	
$V_{T(min)} = 128$ V	
$V_N = 90$ V	

(*c*) $V_{0(min)}$ and $V_{0(max)}$ lie between $V_{T(max)}$ and V_{IG}. Hence no voltage multiplication is necessary ($s = 1$).

(*d*) As no voltage multiplication is needed, it is convenient to use a bridge rectifier to give full-wave rectification ($u = 1$).

(*e*) $$\mathrm{m} = \frac{V_{0(max)}}{V_{0(min)}} = \mathbf{2{\cdot}0}$$

(*f*) Using one tube only ($n = 1$),

$$q' = \frac{1}{n} \cdot \frac{I_{0(av)\,(max)}}{I_{K(av)\,(max)}} = \frac{1}{1} \cdot \frac{10}{25} = \mathbf{0{\cdot}4}$$

(*g*) $$r = \tfrac{1}{2}(r_1 + r_2) = \tfrac{1}{2}(0{\cdot}06 + 0{\cdot}10) = \mathbf{0{\cdot}08}$$

H

(*h*) $$k' = \frac{m(1-w) - q'(1+w)}{(1-r)(1-w) - q'(1+r)(1+w)} \cdot \frac{1}{\sqrt{2}}$$

$$= \frac{2{\cdot}0 \times 0{\cdot}95 - 0{\cdot}40 \times 1{\cdot}05}{0{\cdot}92 \times 0{\cdot}95 - 0{\cdot}40 \times 1{\cdot}08 \times 1{\cdot}05} \cdot \frac{1}{\sqrt{2}} = \mathbf{2{\cdot}48}$$

(*j*) From Fig. 5.7, $p = \mathbf{1{\cdot}93}$

(*k*) $$u \cdot \frac{I_{K(pk)\,(max)}}{I_{K(av)\,(max)}} = 1 \times \frac{100}{25} = 4 \text{ (which is } > p)$$

(Note that this condition would still be satisfied if $u = \frac{1}{2}$. Hence a half-wave rectifier could be used in this case.)

(*l*) $$V_m = \frac{k'}{s} \cdot V_{0(min)} \Big/ \left(1 + \frac{r_1 - r_2}{2}\right)$$

$$= \frac{2{\cdot}48}{1} \times 145 \Big/ \left(1 + \frac{0{\cdot}06 - 0{\cdot}10}{2}\right) = \mathbf{367\ V\ r.m.s.}$$

(*m*) $$R_1 = \frac{u}{n} \cdot \frac{(1 + r_1)\,.\,V_m\sqrt{2} - V_{0(min)}/s}{p\,.\,I_{K(av)\,(max)}}$$

$$= \frac{1}{1} \cdot \frac{1{\cdot}06 \times 367\sqrt{2} - 145/1}{1{\cdot}93 \times 25} = \mathbf{8{\cdot}4\ k\Omega}$$

(*n*) $$R_4 = u\,.\,\frac{V_{0(min)}/s - V_M}{p\,.\,I_{K(av)\,(max)}} = 1 \cdot \frac{145/1 - 105}{1{\cdot}95 \times 25} = \mathbf{0{\cdot}82\ k\Omega}$$

(*o*) **n** $= \mathbf{1}$ $\therefore$ $R_6 = 0$

$$R_5 = R_4 - R_6 = R_4 = \mathbf{0{\cdot}82\ k\Omega}$$

$$R_7 \approx \frac{132 - 90}{10 \times 0{\cdot}045} = \mathbf{93{\cdot}5\ k\Omega}$$

(*q*) $$R_8 = \left(\frac{V_{0(min)}}{s\,.\,V_{T(max)}} - 1\right) \cdot R_7 \cdot \left(\frac{1-w}{1+w}\right)$$

$$= \left(\frac{145}{137} - 1\right) \times 93{\cdot}5 \times \frac{0{\cdot}95}{1{\cdot}05} = \mathbf{5{\cdot}07\ k\Omega}$$

(*r*) $$R_9 = \frac{1}{n}\left\{\left(\frac{V_{0(max)}}{s\,.\,V_{T(min)}} - 1\right).\,R_7 - R_8\right\} \cdot \left(\frac{1+w}{1-w}\right)$$

$$= \frac{1}{1} \cdot \left\{\left(\frac{290}{128} - 1\right) \times 93{\cdot}5 - 5{\cdot}07\right\} \times \left(\frac{0{\cdot}95}{1{\cdot}05}\right) = \mathbf{102{\cdot}5\ k\Omega}$$

To rationalize values, reduce R_7, R_8, and R_9 by 2·5% each. Thus $R_7 = 91$ kΩ, $R_8 = 4{\cdot}95$ kΩ, $R_9 = 100$ kΩ. Since R_8 is small compared

with R_7 and R_9, its precise value is not critical. Using 5% tolerance components, therefore, the following nominal values will be acceptable:

$$R_7 = 91\ k\Omega$$
$$R_8 = 5\ k\Omega$$
$$R_9 = 100\ k\Omega$$

(*s*) $$C \geqslant \frac{s}{8fR_0} \cdot \left(\frac{2-u}{u}\right)$$

$$= \frac{1}{8 \times 50 \times 30} \times \left(\frac{2-1}{1}\right) = \mathbf{83\ \mu F}$$

(*t*) $$C \geqslant \frac{1}{4f} \cdot \left(\frac{2-u}{u}\right) \cdot \frac{I_{0(max)}}{V_r}$$

$$= \frac{1}{4 \times 50} \times \left(\frac{2-1}{1}\right) \times \frac{10 \times 10^{-3}}{1{\cdot}45} = \mathbf{34{\cdot}6\ \mu F}$$

(*u*) $C > 83$ µF. Hence, remembering that operation (*s*) gives a conservative value for *C*, put **C** = **100µF, ±20%.**

Solution: (Using circuit of Fig. 5.10 with only one tube)

$R_1 = 8{\cdot}4$ kΩ ± 5%	$C = 100$ µF ± 20%
$R_5 = 0{\cdot}82$ kΩ ± 5%	V_1 = CV2434
$R_6 = 0$	$V_m = 367$ V r.m.s.
$R_7 = 91$ kΩ ± 5%	
$R_8 = 5$ kΩ ± 5%	
$R_9 = 100$ kΩ ± 5%	

EXAMPLE 5.3 Multi-tube Trigger-tube Shunt Stabilizer

Design a trigger-tube shunt stabilizer to operate from 50 c/s supply at 240 V r.m.s. (+6%, −10%) and to deliver 0–50 mA at 450 V d.c. ±20 V. Output impedance to be less than 100 Ω. Use 5% tolerance resistors.

(*a*) Setting out design data:

$V_{0(max)} = 470$ V	$I_{0(av)\,(max)} = 50$ mA	$f = 50$ c/s
$V_{0(min)} = 430$ V	$r_1 = 0{\cdot}06$	
$V_{r(max)}$ Not specified.	$r_2 = 0{\cdot}10$	
$R_{0(max)} = 100\ \Omega$	$w = 0{\cdot}05$	

(*b*) Basing design on CV2434, use data given in Example 5.2.

(*c*) $V_{0(min)}$ and $V_{0(max)}$ lie between $2V_{T(max)}$ and $2V_{IG}$. Hence voltage doubling is needed ($s = 2$).

(*d*) For convenience of voltage doubling, use half-wave rectification ($u = \frac{1}{2}$).

(*e*) $\quad m = \dfrac{V_{0(max)}}{V_{0(min)}} = \dfrac{470}{430} = \mathbf{1{\cdot}09}$

(*f*) If two tubes are used in parallel ($n = 2$),

$$q' = \frac{1}{2} \times \frac{50}{25} = 1{\cdot}0$$

But q' must be less than 1·0, hence put $n = 3$,

Then $\quad q' = \dfrac{1}{3} \times \dfrac{50}{25} = \mathbf{0{\cdot}67}$

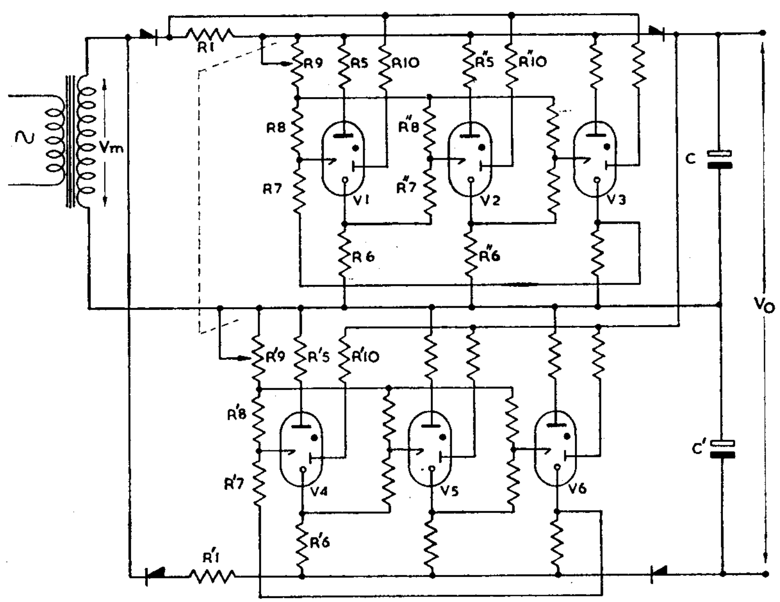

FIG. 5.11. Stabilized supply for 450 V at 0–50 mA (Example 5.3).

(*g*) $\quad r = \frac{1}{2}(0{\cdot}06 + 0{\cdot}10) = \mathbf{0{\cdot}08}$

(*h*) $\quad k' = \dfrac{1{\cdot}09 \times 0{\cdot}95 - 0{\cdot}67 \times 1{\cdot}05}{0{\cdot}92 \times 0{\cdot}95 - 0{\cdot}67 \times 1{\cdot}08 \times 1{\cdot}05} \cdot \dfrac{1}{\sqrt{2}} = \mathbf{2{\cdot}03}$

(*j*) From Fig. 5.7, $p = 1{\cdot}99$.

(*k*) $\quad u \cdot \dfrac{I_{K(pk)\,(max)}}{I_{K(av)\,(max)}} = \dfrac{1}{2} \cdot \dfrac{100}{25} = 2{\cdot}0$ (which is $>p$)

(*l*) $\quad V_m = \dfrac{2{\cdot}03}{2} \times \dfrac{430}{0{\cdot}98} = \mathbf{428\ V\ r.m.s.}$

(*m*) $\quad R_1 = \frac{1}{3} \times \dfrac{1{\cdot}06 \times 428\sqrt{2} - 430/2}{1{\cdot}99 \times 25} = \mathbf{1{\cdot}45\ k\Omega}$

(*n*) $\quad R_4 = \dfrac{1}{2} \times \dfrac{430/2 - 105}{1{\cdot}99 \times 25} = \mathbf{1{\cdot}10\ k\Omega}$

(*o*) **n = 3**

$$\therefore\ R_6 = \frac{(132 - 90) \times 1{\cdot}1}{430/2 - 105} = \mathbf{0{\cdot}42\ k\Omega}$$

$$R_5 = R_4 - R_6 = 1{\cdot}10 - 0{\cdot}42 = \mathbf{0{\cdot}68\ k\Omega}$$

(*p*) Choose

$$R_7 \approx \frac{132 - 90}{10 \times 0{\cdot}045} = \mathbf{93{\cdot}5\ k\Omega}$$

$\therefore$ put $R_7 = \mathbf{100\ k\Omega}$

(*q*) $\quad R_8 = \left(\dfrac{430}{2 \times 137} - 1\right) \times 100 \times \dfrac{0{\cdot}95}{1{\cdot}05} = \mathbf{51{\cdot}5\ k\Omega}$

(*r*) $\quad R_9 = \dfrac{1}{3}\left\{\left(\dfrac{470}{2 \times 128} - 1\right) \times 100 - 51{\cdot}5\right\} \times \dfrac{1{\cdot}05}{0{\cdot}95} = \mathbf{12{\cdot}0\ k\Omega}$

(*s*) $\quad C \geqslant \dfrac{2}{8 \times 50 \times 100} \times \left(\dfrac{2 - \frac{1}{2}}{\frac{1}{2}}\right) = 150\ \mu F$, say $\mathbf{160\ \mu F}$

Solution: (Circuit shown in Fig. 5.11)

$R_1 = 1{\cdot}45\ k\Omega\ \pm 5\%$	$C = 160\ \mu F$
$R_5 = 0{\cdot}68\ k\Omega\ \pm 5\%$	$V_1 - V_6 =$ CV2434
$R_6 = 0{\cdot}42\ k\Omega\ \pm 5\%$	$V_m = 428$ V r.m.s.
$R_7 = 100\ k\Omega\ \pm 5\%$	
$R_8 = 51{\cdot}5\ k\Omega\ \pm 5\%$	
$R_9 = 12{\cdot}0\ k\Omega\ \pm 5\%$	
$R_{10} = 10\ M\Omega\ \pm 20\%$	

Timing Circuits

Trigger tubes are widely used in timing circuits of various kinds. These circuits have been fully investigated and they may be designed with confidence.

Fig. 5.12 illustrates the common basis of all these circuits. A capacitor C is charged through resistor R from a source V_P. After a time t the voltage across C is given by

$$V_C = V_P\,.\,(1 - e^{-t/CR}) \qquad (5.31)$$

When the voltage V_C across C reaches the trigger breakdown potential, V_T, the tube fires to terminate the timed interval. Substituting V_T for V_C in Equation (5.31), the timed interval t is given by

$$t = CR \, . \log_e \left(\frac{V_P}{V_P - V_T} \right) \tag{5.32}$$

In practice, there is an uncertainty δV_T in the value of V_T, and this gives rise to an uncertainty δt in the timed interval. By differentiating Equation (5.32), one obtains

$$\delta t = \frac{CR}{(V_P - V_T)} \, . \, \delta V_T \tag{5.33}$$

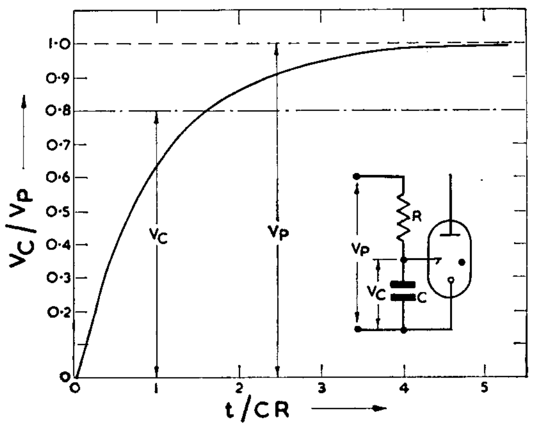

Fig. 5.12. Basis of *RC* timing circuit.

The accuracy of timing, $\delta t/t$, is given by combining Equations (5.32) and (5.33).

$$\frac{\delta t}{t} = \frac{\delta V_T}{(V_P - V_T)} \cdot \frac{1}{\log_e \left(\dfrac{V_P}{V_P - V_T} \right)}$$

$$= \frac{-x}{(1 - x) \, . \log_e (1 - x)} \cdot \frac{\delta V_T}{V_T}$$

$$= \beta \cdot \frac{\delta V_T}{V_T} \tag{5.34}$$

where $x = V_T/V_P$

and $\beta = \dfrac{-x}{(1 - x) \, . \log_e (1 - x)}$

The curve in Fig. 5.13, due to Light [6], shows the relation between the 'error multiplying factor', β, and V_T/V_P. It is seen that when V_T/V_P exceeds 0·8 the accuracy of timing falls rapidly, but if V_T/V_P does not exceed 0·7 a 1% change in V_T produces less than 2% change in t.

In Fig. 5.14, derived from Equation (5.32), t is plotted against V_T/V_P. From this it is seen that t falls abruptly when V_T/V_P is made less than 0·4. Above $V_T/V_P = 0{\cdot}8$, t increases rapidly, but Fig. 5.13 shows this gives poor accuracy. It is thus usual to work with V_T/V_P between 0·4 and 0·8, i.e. $t = 0{\cdot}5CR$ to $1{\cdot}5CR$ approximately. Often it is convenient to put $V_T/V_P = 0{\cdot}63$ so that $t = CR$. The error multiplying factor, β, is the *n* 1·72.

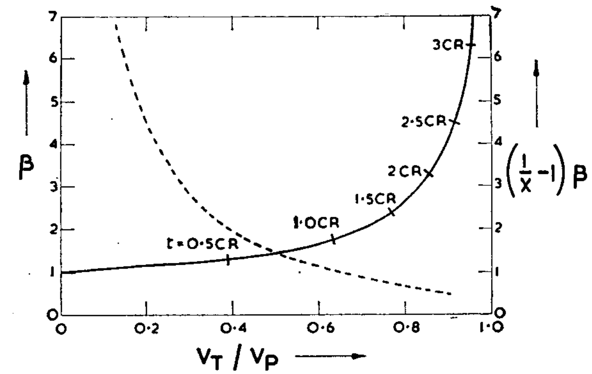

FIG. 5.13. Error multiplying factors for changes in V_P (full line) and V_O (broken line) as functions of V_T/V_P.

As the capacitor voltage approches V_T, a small pre-strike current (p. 68) flows through the trigger electrode. Charging of the capacitor will cease if the pre-strike current I_P exceeds the current which can flow through the charging resistor. Thus the maximum value of R which will produce triggering is given by

$$R_{(max)} = (V_P - V_T)/I_P \tag{5.35}$$

Putting this value into Equation (5.32), the maximum value of t is given as

$$t_{(max)} = C \cdot \frac{(V_P - V_T)}{I_P} \cdot \log_e \left(\frac{V_P}{V_P - V_T} \right)$$

$$= \frac{C \, . \, V_T}{I_P} \cdot \frac{1}{\beta} \tag{5.36}$$

The limiting value of $t_{(max)}$ is thus obtained with the minimum value of β. Fig. 5.13 shows this is $\beta = 1$, obtained when $V_P = \infty$. Even with V_P infinitely large, however, the value of $t_{(max)}$ is only about 2·5 times that obtainable with $V_T/V_P = 0{\cdot}8$. If difficulty is experienced in providing a large value of t, therefore, there is little to be gained by raising

V_P, provided V_P already exceeds $V_T/0{\cdot}8$. A tube should be sought having a lower pre-strike current.

To ensure that the operative value of I_P in Equation (5·36) is the pre-strike current, rather than the transfer current, it may be necessary to use a trigger circuit corresponding to Fig. 4.6 (*d*) rather than Fig. 4.6 (*c*).

Crowther and Potter [8] have studied the various errors arising in a

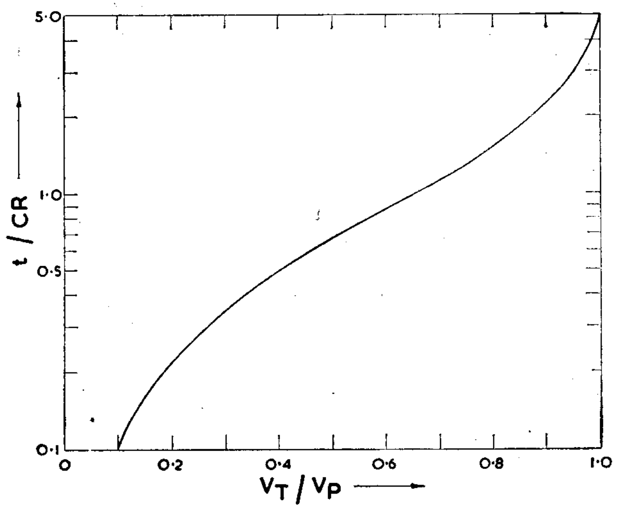

Fig. 5.14. t/CR as a function of V_T/V_P.

timing circuit using the Z803U. They find that for intervals in excess of 10 msec the triggering delay is less than 1%, and this percentage becomes gradually less as t is increased. Over a restricted range of times this delay may be considered part of the timed interval.

If the trigger tube operates a relay the pull-in time of the relay must be considered. This tends to add a constant time, rather than a constant percentage, to the timed interval. If the h.t. voltage falls by only 10%, however, Crowther and Potter point out that the resultant 20% fall in relay current can readily lead to a doubling of the relay pull-in time. For this reason the anode circuit should be designed so that, by making use of the peak current rating of the tube, a relatively large pull-in current is carried by the relay. In this way consistently short pull-in delays are obtained permitting intervals of more than 1 or 2 sec to be timed to an accuracy of better than 1%.

Changes in the values of C and R used for timing produce proportional

errors in the timed intervals. The effect of variations in V_T is shown by Equation (5.34). By differentiating Equation (5.32) with respect to V_P, an expression is obtained for the variation $\delta t'$ due to a variation in V_P.

$$\delta t' = \frac{-CR}{(V_P - V_T)} \cdot \frac{V_T}{V_P} \cdot \delta V_P \qquad (5.36)$$

whence

$$\frac{\delta t'}{t} = -\beta \cdot \frac{\delta V_P}{V_P} \qquad (5.37)$$

Provided the changes in V_T and V_P are sufficiently small for β to be assumed constant, the total variation in t to be expected with given variations of V_T and V_P may be obtained merely by adding the contributions defined separately in Equations (5.34) and (5.37). Since the percentage variations in V_T and V_P are each multiplied by the same factor, β, no great improvement in accuracy can be obtained by reducing the percentage variation of V_T below that of V_P, or vice versa.

The initial voltage on the capacitor may not always be zero. If the initial voltage is V_0 the above analysis holds provided $(V_P - V_0)$ is substituted for V_P and $(V_T - V_0)$ for V_T. Equation (5.32) then becomes

$$t = CR \, . \log_e \frac{(V_P - V_0)}{(V_P - V_T)} \qquad (5.32a)$$

For small variations in V_0,

$$\delta t'' = \frac{-CR}{(V_P - V_0)} \, . \, \delta V_0$$

$$\therefore \quad \frac{\delta t''}{t} = \frac{-1}{(V_P - V_0) \, . \log_e \dfrac{(V_P - V_0)}{(V_P - V_T)}} \, . \, \delta V_0 \qquad (5.38)$$

The corresponding error multiplying factor is a function of both V_0/V_P and V_T/V_P.

A case commonly requiring consideration is that in which V_0 is zero, but it is necessary to determine the error due to the initiation of a new timing cycle before the capacitor has discharged below a residual voltage, δV_0.* Then, since V_0 is zero, Equation (5.38) reduces to

$$\frac{\delta t''}{t} = \frac{-1}{\log_e \dfrac{V_P}{(V_P - V_T)}} \cdot \frac{\delta V_0}{V_P}$$

$$\doteqdot -\left(\frac{1}{x} - 1\right) . \, \beta . \, \frac{\delta V_0}{V_P} \qquad (5.38a)$$

* An initial charge V_0 can also arise due to 'soak' effects in paper dielectric timing capacitors. Plastic film capacitors show negligible soak effect.

The broken line in Fig. 5.13 shows that $\left(\frac{1}{x} - 1\right) . \beta$ decreases as x increases. A value of $V_T/V_P = 0{\cdot}63$ gives a satisfactory compromise between sensitivity to changes in V_T, V_P, and V_0.

EXAMPLE 5.4 Interval Timer

Suppose it is required to design a timer to operate on 240 V a.c. $\pm 10\%$ and to provide over 10,000 hours life a timed interval of 1 sec with the highest accuracy consistent with reasonable economy. The basic circuit of Fig. 5.15 is to be used, and the necessary component values are to be determined.

Choose the CV2434 as a high-stability trigger tube for which it is known that, over 10,000 hours, $\delta V_T/V_T < \pm 2\%$.

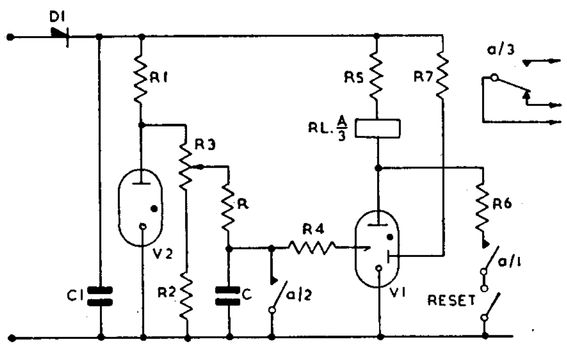

Fig. 5.15. Basic circuit for interval timer.

For this tube, $V_T = 132$ V. Thus if, to keep β low, V_T/V_P is not allowed to exceed 0·8, the minimum value of V_P may be determined,

$$V_{P(\min)} = \frac{V_T}{0{\cdot}8} = 165 \text{ V}$$

The stabilizer, V_2, should therefore have a maintaining voltage of 165 V or more and a stability of $\pm 2\%$ or better through 10,000 hours. The high maintaining voltage calls for two or more tubes operating in series, and only pure metal cathodes will provide the required degree of stability. Two or three CV449 (85A2) voltage reference diodes may therefore be used in series.* To operate these at 6 mA, the series resistance, R_1, must be reduced if more tubes are used in series. In an extreme case the consequent reduction in

* A more economical arrangement would use a single trigger tube as stabilizer in place of the voltage reference tubes. The diode stabilizers are here considered as representing a more conventional design.

stabilization ratio could offset the advantage of a higher effective V_P. A comparison of the two cases is therefore required.

(*a*) With two CV449 tubes in series:

$$V_B 240\sqrt{2} = 340 \text{ V}$$

$$V_P = 170 \text{ V}$$

$$R_1 = \frac{340 - 170}{6} = \mathbf{28\ k\Omega\ approx.}$$

From Equation (3.8a),

$$S \approx \frac{R_S}{R_I} = \frac{28}{0{\cdot}3} = 93$$

Hence a 10% change in V_B produces $10/93 = 0{\cdot}11\%$ change in V_P. This must be added to the 1·5% change in V_P expected through 10,000 hours.

Thus $$\frac{\delta V_P}{V_P} = 1{\cdot}5 + 0{\cdot}11 = 1{\cdot}6\%$$

Now, $$\frac{V_T}{V_P} = \frac{132}{170} = 0{\cdot}78$$

$$\therefore\ \beta = 2{\cdot}4 \text{ (from Fig. 5.13)}$$

$$\therefore\ \frac{\delta t}{t} + \frac{\delta t'}{t} = \beta . \left(\frac{\delta V_T}{V_T} + \frac{\delta V_P}{V_P}\right)$$

$$= 2{\cdot}4 \times (2{\cdot}0 + 1{\cdot}6) = \mathbf{8{\cdot}65\%}$$

(*b*) With three CV449 tubes in series:

$$R_1 = \frac{340 - 255}{6} = \mathbf{14\ k\Omega\ approx.}$$

$$S = \frac{14}{0{\cdot}3} = 47$$

$$\therefore\ \frac{\delta V_P}{V_P} = 1{\cdot}5 + 0{\cdot}21 = 1{\cdot}7\%$$

$$\frac{V_T}{V_P} = \frac{132}{255} = 0{\cdot}52$$

$$\beta = 1{\cdot}5$$

$$\therefore\ \frac{\delta t}{t} + \frac{\delta t'}{t} = 1{\cdot}5 \times (2{\cdot}0 + 1{\cdot}7) = \mathbf{5{\cdot}5\%}$$

There is thus a significant advantage in using three reference tubes in series. Since the worst combination of variations has been considered, the total variation in time interval will normally be much less than 5%.

Determination of other component values may now begin. Timing errors

will increase less than 1% if V_T/V_P is raised to 0·63. At this value, $t = CR$ and it is convenient to put $C = 0{\cdot}1\ \mu F$, $R = 10\ M\Omega$, high-stability components being used for each. To accommodate component tolerances, the precise value of V_P may be adjusted by a potentiometer connected across the uppermost reference tube.

V_1 is normally extinguished by the shunt circuit through contacts $a/1$. If the reset contacts are held open, however, the tube will remain conducting. The design should arrange that the steady current in these circumstances is the recommended value for long tube life (8 mA). The hold-in current of the relay must be less than this. A 5-kΩ P.O. 3000 relay will meet the case, but its pull-in time is long and indefinite at such a low current. The addition of a capacitor C_2 at the junction of relay $A/3$ and R_5 solves this problem. With C_2 initially charged to 240 V, a peak anode current of 28 mA will flow to pull in the relay smartly.

R_5 must be determined before the value of C_2 may be settled. To restrict the anode voltage of V_1 to 240 V before striking, R_8 must be added, as in Fig. 5.16.

$$\frac{R_8}{R_5 + R_8} = \frac{240}{V_B} = \frac{240}{340}$$

Whence $$R_8 = 2{\cdot}4R_5$$

If I_R is the steady current drawn by V_1 after it has struck but with reset contacts held open to prevent quenching, then

$$V_M + I_R R_{LA} = V_B - (I_R + I_8)\,.\,R_5$$

But the current I_8 flowing through R_8 is given by

$$I_8 = \frac{V_M + I_R R_{LA}}{R_8} = \frac{V_M + I_R R_{LA}}{2{\cdot}4R_5}$$

Substituting and solving for R_5,

$$R_5 = \left[V_B - \left(1 + \frac{1}{2{\cdot}4}\right)(V_M + I_R R_{LA})\right]\Big/ I_R$$

Putting I_R at the recommended value of 8 mA and inserting the known values of V_B, V_M, and R_{LA},

$$R_5 = [340 - 1{\cdot}42 \times (105 + 40)]/8 = 16{\cdot}7\ k\Omega$$

Since V_B will fall when V_1 fires and imposes an additional load on the power supply, it is in order to round off R_5 to a slightly lower value, i.e. **15 kΩ** ±5%.

$$R_8 = 2{\cdot}4R_5 = \mathbf{36\ k\Omega} \pm 5\%$$

When the reset contacts are allowed to close, R_6 must draw at least the current previously drawn by V_1. Thus

$$R_{6(max)} = \frac{V_{M(min)}}{I_R} = \frac{100}{8} = 12{\cdot}5\ k\Omega$$

To allow for component tolerances, R_6 should be 0·7 of this value or less. Thus

$$R_6 \approx 0{\cdot}7 \times 12{\cdot}5 \approx \mathbf{8{\cdot}2\ k\Omega} \pm 10\%$$

If one neglects the inductance of the relay the time-constant of the anode circuit is the product of C_2 and the parallel resistances of R_5, R_{LA}, and R_8. Assuming this time-constant is to exceed 100 msec in order to ensure that the pull-in of the relay is effected by a fairly high current, one obtains (working in microfarads, kilohms, and milliseconds),

$$\frac{C_2 R_5\,.\,R_{LA}\,.\,R_8}{R_5 R_{LA} + R_{LA} R_8 + R_8 R_5} > 100$$

$$\therefore\ C_2 > \frac{100 \times (15 \times 5 + 5 \times 36 + 36 \times 15)}{15 \times 5 \times 36} = 29{\cdot}4\ \mu F$$

Hence put $C_2 = \mathbf{32\ \mu F}$

The complete circuit thus becomes as shown in Fig. 5.16. R_1, which was calculated as 14 kΩ, may be rounded off to 15 kΩ. Relations (3.4c) and (3.7) lead to a tolerance of ±20% on this value. P_1 is made 100 k so that the

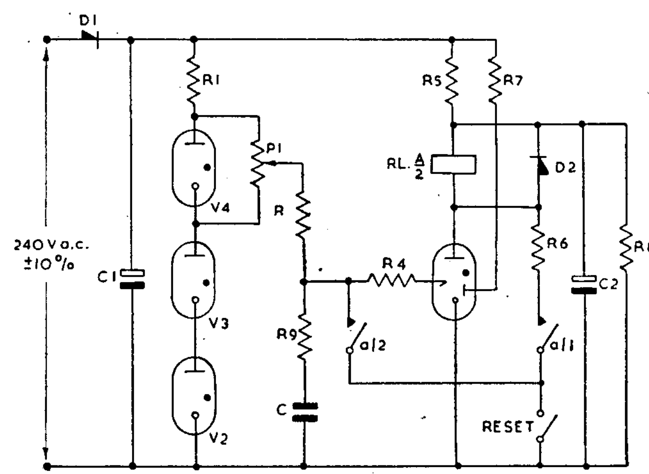

Fig. 5.16. Interval timer designed in Example 5.4.

current in the uppermost reference tube is reduced by less than 1 mA. By adjustment of P_1, V_P may be varied from 170 to 225 V. The ratio V_T/V_P accordingly varies between 0·78 and 0·52, and with it the timed intervals vary from $1{\cdot}5CR$ to $0{\cdot}75CR$. Tolerances in C, R, and V_T can therefore be accommodated by adjustment of P_1.

The arrangement of the discharging contacts, $a/2$, in Fig. 5.15 is unsatisfactory, because if the reset contacts are held open at the end of the interval, V_1 remains conducting with its trigger at ground potential. The consequent

reverse trigger current will produce a substantial change in V_T. In Fig. 5.16 this weakness is eliminated.

Also in Fig. 5.16 a small resistance, R_9, is shown in series with C to limit the discharging current through contacts $a/2$. If $R_9 \ll R$ the time interval is affected to a negligible extent. Here R_9 could be 470 Ω, so that the peak discharge current, V_N/R_9, is about 190 mA.

Throughout the design it has been assumed C_1 is sufficiently large for V_B to approximate the peak supply voltage, despite a current drain of 6 mA through R_1 and 6·7 mA through R_5 and R_8. A value of 16 μF or more meets this requirement.

From the tube manufacturer's data book, R_4 should be 5·6 kΩ.

The final design is thus:

R = 10 MΩ ±5%	R_7 = 10 MΩ ±20%	C = 0·1 μF ±20%
R_1 = 15 kΩ ±20%	R_8 = 36 kΩ ±5%	C_1 = 16 μF ±20%
R_4 = 5·6 kΩ ±20%	R_9 = 470 Ω ±20%	C_2 = 32 μF ±20%
R_5 = 15 kΩ ±5%	R_{LA} = 5 kΩ	V_1 = CV2434 (Z803U)
R_6 = 8·2 kΩ ±10%	P_1 = 100 kΩ ±20%	V_2, V_3, V_4 = CV449 (85A2)

Power Law Timers

There are several devices which, when operated from a variable voltage supply, V_m, provide an energy output proportional to $V_m{}^n$. In a welder or an oven the power is proportional to $V_m{}^2$. The blue-violet light output from a tungsten-filament lamp is proportional to $V_m{}^5$, while the green and red light output correspond to intermediate values of n. Timers designed for use with welders or in photography may therefore be of greatest value if designed to produce time intervals proportional to $1/V_m{}^n$.

Suitable circuits have indeed been developed [7, 9], and these prove simple to design provided the appropriate value of n is known. Not only does the compensated timer eliminate the need for a supply voltage stabilizer but it also proves more economical than a simple timer of comparable accuracy.

If the power output P of the device is expressed as $k \,.\, V_m{}^n$, then

$$\delta P = nk \,.\, V_m{}^{(n-1)} \,.\, \delta V_m \tag{5.39}$$

whence,

$$\frac{\delta P}{P} = n \,.\, \frac{\delta V_m}{V_m} \tag{5.40}$$

During a timed interval the energy output is $P \,.\, t$. Hence the timer is compensated if

$$\frac{\delta P}{P} = -\frac{\delta t}{t}$$

From Equation (5.37), this gives

$$\beta \,.\, \frac{\delta V_P}{V_P} = n \,.\, \frac{\delta V_m}{V_m} \tag{5.41}$$

Thus, if V_P is made proportional to V_m the timer will be exactly compensated when $\beta = n$. In this type of circuit, therefore, the error-multiplying factor associated with changes in V_T is necessarily equal to n when precise compensation is provided. Thus, from Equation (5.34),

$$\frac{\delta(P \,.\, t)}{(P \,.\, t)} = n \,.\, \frac{\delta V_T}{V_T} \tag{5.42}$$

With a high value of n (as in a photographic timer for use with blue-sensitive materials) it is thus necessary to use a tube providing a highly stable value of V_T.

Inspection of Fig. 5.13 shows that for values of β and n greater

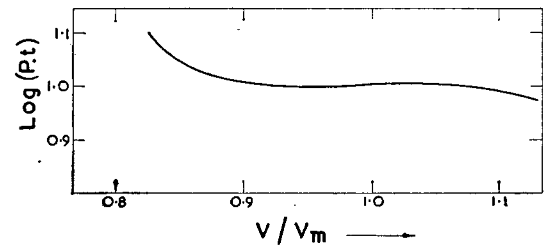

Fig. 5.17. Characteristic of Power × Time provided by more refined type of power-law timer.

than 2, β changes rapidly with V_T/V_P. Consequently, good compensation is provided over only a small range of supply voltages. The timer is under-compensated if V_P increases and over-compensated if V_P falls. When $n = 5$ the required value of $\beta = 5$ is reached at $V_T/V_P = 0{\cdot}93$. If V_P falls by 7·5% $V_T/V_P = 1{\cdot}0$ and the timed interval does not terminate.

For applications requiring high values of β a more sophisticated arrangement has been devised [9]. This comprises applying to the trigger a potential which is a combination of an exponential and a bias potential, each proportional to V_m. By this technique, a characteristic of $P \,.\, t$ against V_m can be obtained which exhibits a plateau as shown in Fig. 5.17.

The simple arrangement shown in Fig. 5.18 provides a bias potential λV_m on which is superimposed the exponential charging of C from the

charging potential αV_m. Effectively the trigger breakdown potential is reduced by the bias λV_m. Hence Equation (5.32) becomes

$$t = CR \, . \log_e \frac{\alpha V_m}{\alpha V_m - (V_T - \lambda V_m)} \tag{5.32a}$$

In a timed interval the energy output is thus given by

$$P \, . \, t = CR \, . \, k V_m{}^n \, . \log_e \frac{\alpha V_m}{(\alpha + \lambda) V_m - V_T} \tag{5.43}$$

Now at the point of inflexion of the characteristic shown in Fig. 5.17, both $\mathrm{d}(P \, . \, t)/\mathrm{d}V_m$ and $\mathrm{d}^2(P \, . \, t)/\mathrm{d}V_m{}^2$ become zero.

Differentiating Equation (5.43) and putting $\mathrm{d}(P \, . \, t)/\mathrm{d}V_m = 0$,

$$\frac{V_T}{(\alpha + \lambda) V_m - V_T} = n \, . \log_e \frac{\alpha V_m}{(\alpha + \lambda) V_m - V_T} = 0 \tag{5.44}$$

Differentiating a second time and putting $\mathrm{d}^2(P \, . \, t)/\mathrm{d}V_m{}^2 = 0$, one obtains,

$$\frac{(\alpha + \lambda)}{(\alpha + \lambda) V_m - V_T} = \frac{n}{V_m} \tag{5.45}$$

and thus,

$$(\alpha + \lambda) V_m = \frac{n}{n-1} \cdot V_T \tag{5.46}$$

Substituting in Equation (5.44),

$$\frac{1}{\left(\frac{n}{n-1}\right) - 1} - n \, . \log_e \left[\frac{\alpha}{\left(\frac{n}{n-1}\right) - 1} \cdot \frac{V_m}{V_T} \right] = 0$$

$$\therefore \; \log_e \left[(n-1) \, . \, \alpha \frac{V_m}{V_T} \right] = \frac{n-1}{n}$$

whence

$$\alpha = \frac{1}{n-1} \cdot \frac{V_T}{V_m} \cdot e^{\left(\frac{n-1}{n}\right)} \tag{5.47}$$

From Equation (5.46),

$$\lambda = \frac{n}{n-1} \cdot \frac{V_T}{V_m} - \alpha$$

$$\therefore \; \lambda = \frac{n - e^{\left(\frac{n-1}{n}\right)}}{n-1} \cdot \frac{V_T}{V_m} \tag{5.48}$$

Thus

$$\frac{\lambda}{\alpha} = n \cdot e^{\left(\frac{1-n}{n}\right)} - 1 \tag{5.49}$$

The required ratio of λ/α thus proves to be a function of n alone and is independent of V_T/V_m. R_1 and R_2 are therefore proportioned to give

the required ratio λ/α and R_3 is adjusted to suit the particular value of V_T/V_M.

From Equation (5.32a),

$$t = -CR \, . \log_e \left[1 + \frac{\lambda}{\alpha} - \frac{V_T}{\alpha V_m} \right]$$

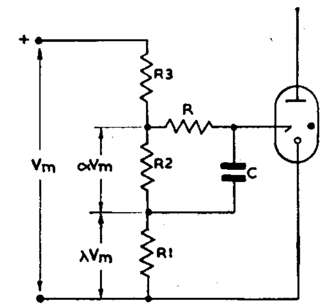

Fig. 5.18. Basic circuit of power-law timer.

Substituting from Equations (5.47) and (5.49),

$$t = -CR \, . \log_e \left[n \, . \, e^{\left(\frac{1-n}{n}\right)} - (n-1) \, . \, e^{\left(\frac{1-n}{n}\right)} \right]$$

$$= -CR \, . \log_e e^{\left(\frac{1-n}{n}\right)}$$

$$\therefore \; t = \left(\frac{n-1}{n}\right) . \, CR \tag{5.50}$$

Note that even with high values of n, $t < CR$.

This value for t corresponds to a design in which the point of inflexion in the characteristic shown in Fig. 5.17 occurs at the nominal supply voltage. In such a case, however, the timer is compensated for a wider range of supply variations above the nominal value than below it. Generally the opposite condition is required. This is readily obtained by making the point of inflexion correspond to about 95% of the nominal supply voltage. All the foregoing relations remain valid except that t is now given by

$$t = 0{\cdot}95^n \, . \left(\frac{n-1}{n}\right) . \, CR \tag{5.50a}$$

When $n = 5$, this gives $t = 0{\cdot}62CR$.

It has been proposed in the past [9] to use half-wave rectified a.c. for the supply V_m in Fig. 5.18. This makes it possible to use large values of

R with a trigger tube in which the trigger–cathode gap has a characteristic apt to provide a stable corona discharge. The pulsating voltage developed across R_1 is transmitted to the trigger by C. As a result, trigger current flows only on the positive peaks, and the requisite pre-strike current can be reached for a much smaller mean current through R. With high values of n, however, this technique usually leads to a design in which reverse trigger current is drawn after striking. It is thus preferable to use smoothed d.c. for V_m in Fig. 5.18 and to choose a stable trigger tube with a negligible corona region in its characteristic.

EXAMPLE 5.5 Photographic Power Law Timer

Suppose a photographic exposure timer is required to control a printer using blue-sensitive materials. The timer is to operate from 50 c/s a.c. supplies of nominal voltages in the range 200–250 V r.m.s. and is to provide nominal time intervals of 1–30 sec compensated according to a fifth-power law for supply variations of at least —15% to +10%.

To meet the wide range of compensation, it is necessary to use the more refined type of circuit shown in Fig. 5.18. Since a fifth-power law is required, $n = 5$.

From Equation (5.50a),

$$CR = t_{\max} \cdot \left(\frac{n}{n-1}\right) \cdot \frac{1}{0{\cdot}95^n}$$

$$= 30 \times \frac{5}{4} \times \frac{1}{0{\cdot}78} = \mathbf{48}\ \text{secs}$$

In the circuit of Fig. 5.19 it is necessary to restrict the value of R so that the back resistance of D_2 shall be negligible. This leads to a rather large value of C. A reasonable compromise is thus $R = 5\ \text{M}\Omega$, $C = 10\ \mu\text{F}$ (paper dielectric). The arrangement shown allows a single contact $a/1$ to provide a hold-in circuit for the relay and (via D_2) a discharging path for the timing capacitor. More obvious arrangements using two reset contacts can cause reverse trigger current to damage the tube should one contact fail to operate.

The resistor chain, R_1, R_2, R_3, is now evaluated as follows:

$$\frac{R_1}{R_2} = \frac{\lambda}{\alpha} = n \cdot e^{\left(\frac{1-n}{n}\right)} - 1 \qquad (5.49)$$

$$= 5 \times 2{\cdot}718^{-0{\cdot}8} - 1 = 1{\cdot}25$$

For any nominal value of V_m, R_4 is adjusted so that $t = 0{\cdot}62CR$. In this condition, the voltage across C_2 is always $(\alpha + \lambda)V_m/0{\cdot}95$.

$$\alpha V_m = \frac{1}{n-1} \cdot e^{\left(\frac{n-1}{n}\right)} \cdot V_T \qquad (5.47a)$$

Hence, if the CV2434 tube is used,

$$\alpha V_m = 0{\cdot}25 \times 2{\cdot}718^{0{\cdot}8} \times 132 = \mathbf{73{\cdot}5\ V}$$

$$\therefore\ (\alpha + \lambda)V_m = \left(1 + \frac{\lambda}{\alpha}\right) \cdot \alpha V_m = 2{\cdot}25 \times 73{\cdot}5 = \mathbf{165\ V}$$

Hence the voltage across C_2 is $165/0{\cdot}95 = 174$ V when the supply is at its nominal voltage (i.e. $t = 0{\cdot}62CR$).

When the supply voltage falls below the nominal value the voltage across

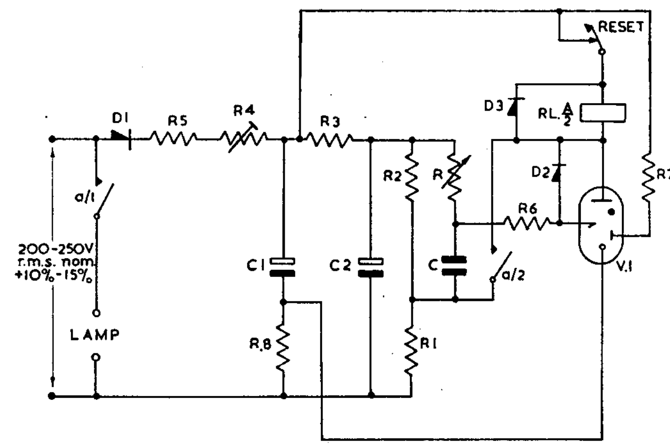

Fig. 5.19. Complete circuit of photographic power-law timer.

C_2 falls below the 170 V which is the recommended minimum for the anode circuit. Accordingly, the anode circuit must be returned to the high-potential side of R_3.

To ensure that $V_A \geqslant 170$ V even when the supply falls to 85% of its nominal voltage,

$$\frac{R_1 + R_2 + R_3}{R_1 + R_2} \times 174 \times 0{\cdot}85 \geqslant 170$$

Remembering $R_1/R_2 = 1{\cdot}25$,

$$R_3 \geqslant 0{\cdot}12R_1$$

To avoid anode–cathode breakdown, $V_A \leqslant 290$ V when the supply voltage rises to 110% of nominal. Thus,

$$\frac{R_1 + R_2 + R_3}{R_1 + R_2} \times 174 \times 1{\cdot}10 \leqslant 290$$

$$\therefore\ R_3 \leqslant 0{\cdot}94R_1$$

Hence R_3 must lie between $0{\cdot}12R_1$ and $0{\cdot}94R_1$. To accommodate a wide tolerance on the value of R_3, a value is chosen well within this range. A more positive relay action is obtained by keeping R_3 towards the upper limit of the range.

The hold-on circuit through $a/1$ and R_1 dictates that $R_1 \lesssim 3{\cdot}3$ kΩ. Thus

$$R_1 = 3{\cdot}3\text{ k}\Omega \pm 2\%$$
$$R_2 = 2{\cdot}7\text{ k}\Omega \pm 2\%$$
$$R_3 = 1{\cdot}8\text{ k}\Omega \pm 5\%$$

Suitable values of R_4 and R_5 may be derived by assuming negligible ripple in the voltage V_B appearing across C_1. Evaluation of R_4 and R_5 then follows much the same course as the determination of R_1 in the trigger-tube voltage stabilizer (p. 102 *et seq*). R_8 is initially assumed to be zero.

If $V_T = 132$ V, then for any nominal supply voltage,

$$V_B = \frac{R_1 + R_2 + R_3}{R_1 + R_2} \times 174 = \frac{7{\cdot}8}{6{\cdot}0} \times 174 = 226\text{ V}$$

If the value of V_T is on the lower tolerance limit ($V_T = 128$ V), however,

$$V_{B(min)} = 226 \times \frac{128}{132} = 219\text{ V}$$

The highest nominal supply voltage, $V_{m(max)}$, is 250 V r.m.s. Thus,

$$k_{(max)} = \frac{V_{m(max)}\sqrt{2}}{V_{B(min)}} = \frac{250\sqrt{2}}{219} = 1{\cdot}61$$

From Fig. 5.7, the corresponding value of p is 2·12.

Now, with $V_T = 128$ V,

$$I_{(av)\,(min)} = \frac{V_{B(min)}}{R_1 + R_2 + R_3} = \frac{219}{7{\cdot}8} = 28{\cdot}1\text{ mA}$$

Half-wave rectification is used, therefore $u = \frac{1}{2}$.

$$I_{(pk)\,(min)} = \frac{p}{u} \cdot I_{(av)\,(min)} = 2{\cdot}12 \times 2 \times 28{\cdot}1 = 119\text{ mA}$$

$$\therefore\ (R_4 + R_5)_{(max)} \geqslant \frac{V_m\sqrt{2} - V_{B(min)}}{I_{(pk)\,(min)}} = \frac{250\sqrt{2} - 219}{119} = \mathbf{1{\cdot}14\ k\Omega}$$

At the opposite extreme, $V_T = 137$ V.

Then, $$V_{B(max)} = 226 \times \frac{137}{132} = 234\text{ V}$$

and $$I_{(av)\,(max)} = \frac{V_{B(max)}}{R_1 + R_2 + R_3} = \frac{234}{7{\cdot}8} = 30{\cdot}1\text{ mA}$$

When $$V_m = 200\text{ V r.m.s.},$$

$$k_{(min)} = \frac{200\sqrt{2}}{234} = 1{\cdot}21$$

From Fig. 5.7, the corresponding value of $p = 2{\cdot}51$.

$$\therefore\ I_{(pk)\,(max)} = 2{\cdot}51 \times 2 \times 30{\cdot}1 = 151\text{ mA}$$

$$\therefore\ R_5 \leqslant \frac{200\sqrt{2} - 234}{151} = \mathbf{0{\cdot}325\ k\Omega}$$

Hence, if $R_8 = 0$, $R_5 = 220$ Ω, ±10%, $R_4 = 1$ kΩ, ±5%.

The voltage $\lambda V_m/0{\cdot}95$ across R_1 is normally given by:

$$\frac{\lambda V_m}{0{\cdot}95} = \frac{\lambda}{\alpha} \cdot \frac{\alpha V_m}{0{\cdot}95} = 1{\cdot}25 \times \frac{73{\cdot}5}{0{\cdot}95} = 97\text{ V}$$

As this is less than $V_{M(min)}$, V_1 will normally extinguish when $a/1$ closes. The extreme case must be checked, however. If $V_{T(max)} = 137$ V, and the supply voltage is 10% high, then the voltage, V_{R1}, across R_1, is given by:

$$V_{R1} = 97 \times \frac{137}{132} \times 1{\cdot}10 = 111\text{ V}$$

This exceeds $V_{M(min)}$ by 11 V, and hence some tubes will not always quench unless the resistance R_8 is included as shown in Fig. 5.19. Ripple current passing through C_1 then causes the cathode of V_1 to be pulsed positively. If these pulses exceed 11 V, V_1 will deionize irrespective of tube and supply variations within the design tolerances. Since the corresponding value of $I_{(pk)}$ has been calculated to be 151 mA, reliable operation is assured if $R_8 = 120$ Ω. This will require a corresponding reduction in the value of R_5, since R_4, R_5, and R_8 are all in series with the ripple current flowing through C_1. Thus suitable values would be $R_5 = 100$ Ω, ±5%, $R_8 = 120$ Ω, ±5%.

Since these three resistors all carry strongly pulsating currents, their wattage ratings must be considerably in excess of those required merely on consideration of the mean d.c. currents flowing.

Solution

$R = 150$ kΩ – 5 MΩ	$C = 10$ μF (paper)
$R_1 = 3{\cdot}3$ kΩ ±2% (5-watt)	$C_1 = 16$ μF 350 V
$R_2 = 2{\cdot}7$ kΩ ±2% (2-watt)	$C_2 = 16$ μF, 250 V
$R_3 = 1{\cdot}8$ kΩ ±5% (2-watt)	$D_1 = 50$ mA, 700 V P.I.V.
$R_4 = 1$ kΩ variable (5-watt)	$D_2 = 25$ mA, 200 V P.I.V.
$R_5 = 100$ Ω ±5% (2-watt)	(>100 MΩ reverse resistance)
$R_6 = 6{\cdot}8$ kΩ ±20%	$D_3 = 25$ mA, 200 V P.I.V.
$R_7 = 10$ MΩ ±20%	$V_1 = $ CV2434
$R_8 = 120$ Ω ±5% (1-watt)	
$R_{LA} = 5$ kΩ	

Logic Circuits

It was noted in Chapter Four that pulse-plus-bias triggering provides an 'AND' gate. This particular type of gate is better described as

'bias-before-pulse'; this is indicated by the waveforms in Fig. 5.20 (*a*). Other 'AND' gates are free from this restriction. In Fig. 5.20 (*b*) coincidence of a positive trigger pulse and negative cathode pulse produces triggering if their combined amplitudes exceed V_T. The output must

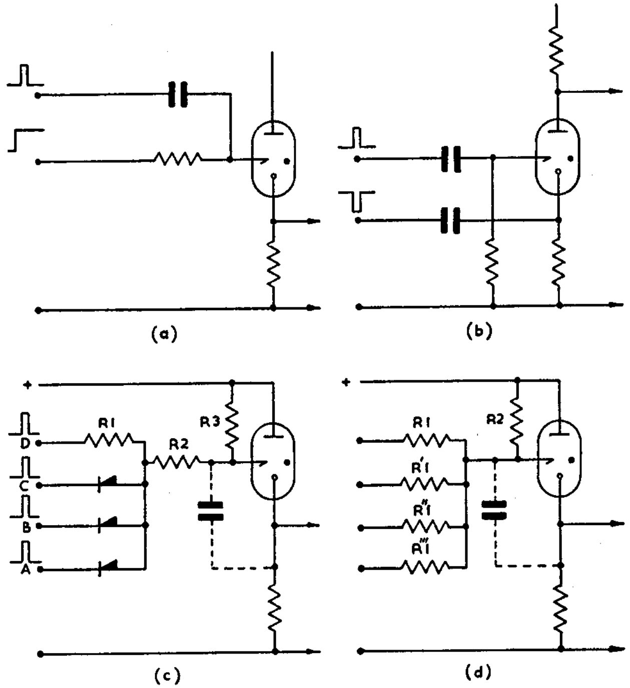

Fig. 5.20. Trigger tube gates: (*a*) Bias-before-pulse 'AND' gate, (*b*) pulse-plus-pulse 'AND' gate, (*c*) diode logic (voltage transfer) 'AND' gate, (*d*) '*m* out of *n*' gate.

here be taken from the anode. Figs. 5.20 (*a*) and (*b*) may be combined to provide a three-input gate.

Any number of inputs may be used with the circuit of Fig. 5.20 (*c*). It may be regarded either as a form of voltage-transfer circuit (p. 75) or as a diode gate followed by a trigger tube as power amplifier. This is a particularly useful circuit, giving an output at anode and/or cathode. If n inputs are used only $(n - 1)$ diodes are required, R_1 taking the place of the nth diode.

So that the back resistance, R_D, of the diodes may be neglected, it is necessary that

$$R_1 \ll R_D/n$$

Also, so that the maximum current, I_D, of the diode shall not be exceeded,

$$R_1 > V_{in}/I_D$$

R_1 is therefore chosen conveniently between these limits.

Now, with inputs at A, B, and C, but not D, the tube will not fire provided

$$\frac{(R_1 + R_2)\,.\,(1 + w)}{(R_1 + R_2)(1 + w) + R_3(1 - w)} \cdot V_{B(max)} < V_{T(min)}$$

i.e.

$$\frac{R_3}{R_1 + R_2} \cdot \frac{1 - w}{1 + w} > \frac{V_{B(max)} - V_{T(min)}}{V_{T(min)}} \qquad (5.51)$$

where w is the fractional tolerance on resistors R_1, R_2, and R_3.

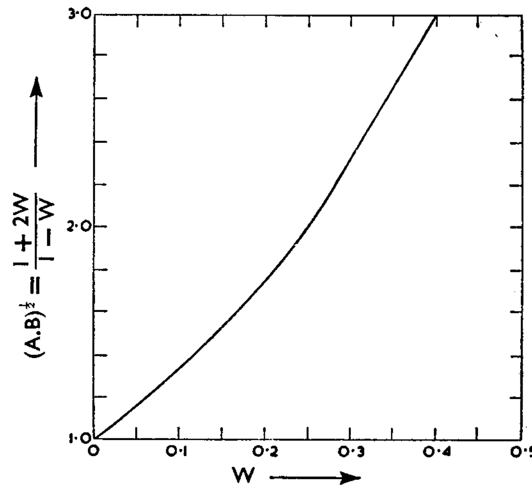

Fig. 5.21. Curve for determination of resistor tolerance, w, in diode logic 'AND' gate.

Putting $R_2/R_1 \geqslant 1/w$, Relation (5.51) gives

$$\frac{R_3}{R_2} \cdot \frac{1 - w}{1 + 2w} > \frac{V_{B(max)} - V_{T(min)}}{V_{T(min)}} \qquad (5.51a)$$

With all inputs energized, the tube is certain to fire and give the 'AND' indication only if

$$\frac{R_2(1 - w)}{R_2(1 - w) + R_3(1 + w)} \cdot (V_{B(min)} - V_{in(min)}) > V_{T(max)} - V_{in(min)}$$

i.e.

$$\frac{R_3}{R_2} \cdot \frac{1 + w}{1 - w} < \frac{V_{B(min)} - V_{T(max)}}{V_{T(max)} - V_{in(min)}} \qquad (5.52)$$

where $V_{in(min)}$ is the minimum value of the nominal input voltage V_{in}.

To provide a small over-voltage on the trigger so that rapid triggering is assured, the term ($V_{T(max)} - V_{in(min)}$) will be increased by a factor $(1 + w)$. Thus Relation (5.52) becomes

$$\frac{R_3}{R_2} \cdot \frac{(1+2w)}{(1-w)} < \frac{V_{B(min)} - V_{T(max)}}{V_{T(max)} - V_{in(min)}} \qquad (5.52a)$$

For a given value of $V_{in(min)}$, the maximum value of w is determined by equating the values of R_3/R_2 given by the limiting conditions in Relations (5.51a) and (5.52a). Hence

$$\left(\frac{1+2w}{1-w}\right)^2 = \frac{V_{B(min)} - V_{T(max)}}{V_{T(max)} - V_{in(min)}} \cdot \frac{V_{T(min)}}{V_{B(max)} - V_{T(min)}} = \text{A.B} \qquad (5.53)$$

The maximum permissible value of w may be read from the curve of $(\text{A.B})^{\frac{1}{2}}$ against w in Fig. 5.21.

Substituting in Relation (5.51a), the corresponding value of R_3/R_2 is then given by,

$$\frac{R_3}{R_2} = \left(\frac{V_{B(min)} - V_{T(max)}}{V_{T(max)} - V_{in(min)}} \cdot \frac{V_{B(max)} - V_{T(min)}}{V_{T(min)}}\right)^{\frac{1}{2}} = (\text{A/B})^{\frac{1}{2}} \qquad (5.54)$$

Design Procedure for Diode Logic 'AND' Gate

(*a*) Set out the following data for the selected tube:
Anode breakdown potential, V_{IG}
Minimum recommended anode potential, $V_{A(min)}$
Maximum trigger striking potential, $V_{T(max)}$
Minimum trigger striking potential, $V_{T(min)}$

(*b*) Set out the following additional data:
Maximum input signal, $V_{in(max)}$
Minimum input signal to operate logic, $V_{in(min)}$
Maximum h.t. voltage, $V_{B(max)}$ ($< V_{IG}$)
Minimum h.t. voltage, $V_{B(min)}$ ($> V_{A(min)}$)
Number of inputs to gate, n

(*c*) Choose diode to have P.I.V. $> V_{in(max)}$

(*d*) Set out data on diode:
Minimum back resistance, R_D
Maximum forward current, I_D

(*e*) Evaluate,

$$\text{A} = \frac{V_{B(min)} - V_{T(max)}}{V_{T(max)} - V_{in(min)}}$$

(*f*) Evaluate,

$$\text{B} = \frac{V_{T(min)}}{V_{B(max)} - V_{T(min)}}$$

(*g*) Evaluate $(\text{A.B})^{\frac{1}{2}}$ and use Fig. 5.21 to determine fractional resistor tolerance, w. (If this is unacceptably small, either increase $V_{in(min)}$, or choose a tube with a lower value of V_T.)

(*h*) Evaluate $(\text{A/B})^{\frac{1}{2}}$.

(*j*) Choose

$$\frac{2V_{in}}{I_D} < R_1 \ll \frac{R_D}{n}$$

(*k*) Choose

$$\frac{R_1}{w} < R_2 > \frac{2V_{B(max)}}{I_D[1 + (\text{A/B})^{\frac{1}{2}}]}$$

(*l*) Evaluate

$$R_3 = (\text{A/B})^{\frac{1}{2}} \,.\, R_2 \qquad (5.54a)$$

EXAMPLE 5.6 Diode Logic 'AND' Gate

Design a 5-input 'AND' gate for inputs of 100 V ± 20%. Use type CV2434 trigger tube in circuit of Fig. 5.20 (*c*).

(*a*) $V_{IG} = 290$ V $V_{T(max)} = 137$ V
$V_{A(min)} = 170$ V $V_{T(min)} = 128$ V

(*b*) $V_{in(max)} = 120$ V Put $V_{B(max)} = 275$ V
$V_{in(min)} = 80$ V $V_{B(min)} = 225$ V
$n = 5$

(*c*) A suitable diode is the OA202.

(*d*) For the OA202, $R_D = 15$ MΩ (at 125° C)
$I_D = 48$ mA (at 125° C)

(*e*) $$\text{A} = \frac{225 - 137}{137 - 80} = \mathbf{1{\cdot}54}$$

(*f*) $$\text{B} = \frac{128}{275 - 128} = \mathbf{0{\cdot}87}$$

(*g*) $$(\text{A.B})^{\frac{1}{2}} = (1{\cdot}54 \times 0{\cdot}87)^{\frac{1}{2}} = \mathbf{1{\cdot}16}$$

From Fig. 5.21, $w_{(max)} = \mathbf{0{\cdot}05}$

(*h*) $$(\text{A/B})^{\frac{1}{2}} = (1{\cdot}54/0{\cdot}84)^{\frac{1}{2}} = \mathbf{1{\cdot}33}$$

(*j*) $$\frac{2V_{in}}{I_D} = \frac{2 \times 100}{48} = 4{\cdot}2 \text{ k}\Omega$$

$$\frac{R_D}{n} = \frac{15{,}000}{5} = 3{,}000 \text{ k}\Omega = 3 \text{ M}\Omega$$

Hence put $R_1 = \mathbf{33\ k\Omega}$

(*k*) $$R_2 > \frac{2 \times 275}{48[1 + 1{\cdot}33]} = 4{\cdot}92\ \text{k}\Omega$$

and, $$R_2 > R_1/w = 33/0{\cdot}05 = 660\ \text{k}\Omega$$

Hence put $$R_2 = \mathbf{680\ k\Omega}$$

(*l*) $$R_3 = 1{\cdot}33R_2 = \mathbf{910\ k\Omega}$$

Solution:

$R_1 = 33$ kΩ ±5% $D_1 = D_2 = D_3 = D_4 =$ OA202
$R_2 = 680$ kΩ ±5% $V_B = 250$ V ± 25 V
$R_3 = 910$ kΩ ±5%

'm out of n' Gate Circuit

Fig. 5.20 (*d*) represents a simple resistor-net adding circuit which may be used to provide either an 'AND' gate or a gate giving an output when any m (or more) of the n inputs are energized. This latter is a logic which would otherwise call for a multiplicity of 'AND' gates, each feeding into an 'OR' gate. If close tolerances are to be avoided, n should not exceed about 5.

For reliable operation, the difference in voltage at trigger A with $(m - 1)$ and m inputs energized must be greater than the variation due to resistor tolerances. With m inputs energized, and R_2 omitted,

$$V_{A(m)} = V_{in} \cdot \frac{m}{n} \tag{5.55}$$

With $(m - 1)$ inputs energized,

$$V_{A(m-1)} = V_{in} \cdot \frac{(m-1)}{n} \tag{5.56}$$

The maximum variation in V_A with resistor tolerance w is $(1 + w) : (1 - w)$. Hence for reliable operation it is necessary that

$$V_{in} \cdot \frac{(m-1)}{n} \cdot \frac{(1+w)}{(1-w)} < V_{in} \cdot \frac{m}{n} \cdot \frac{(1-w)}{(1+w)} \tag{5.57}$$

i.e. $$\left(\frac{1+w}{1-w}\right)^2 < \frac{m}{m-1} \tag{5.58}$$

If no other factors had to be considered one could deduce from Relation (5.58) the following theoretical maximum values of m:

Resistor tolerance	1%	2%	5%	10%	20%
$w =$	0·01	0·02	0·05	0·10	0·20
Maximum value of m	26	14	5	3	1

In practice, additional tolerances further reduce the permissible value of m. Account must be taken of the fractional tolerance, $\pm v$, on the input voltage, V_{in}. Thus

$$V_{in} \cdot (1 + v) \cdot \frac{(m-1)}{n} \cdot \frac{(1+w)}{(1-w)} < V_{in} \cdot (1 - v) \cdot \frac{m}{n} \cdot \frac{(1-w)}{(1+w)} \tag{5.57a}$$

i.e. $$\left(\frac{1+v}{1-v}\right) \cdot \left(\frac{1+w}{1-w}\right)^2 < \frac{m}{m-1} \tag{5.58a}$$

For every 2% which is added to the tolerance of V_{in}, therefore, the resistor tolerance must be reduced by about 1% if the maximum permissible value of m is not to be reduced.

Allowance must be made also for the fractional tolerance, s, on the trigger striking potential, V_T. At best this leads to

$$V_{in} \cdot (1 + v) \cdot \frac{(m-1)}{n} \cdot \frac{(1+w)}{(1-w)} + 2sV_T\left(\frac{R_2 + R_1/n}{R_2}\right) < V_{in} \cdot (1 - v) \cdot \frac{m}{n} \cdot \frac{(1-w)}{(1+w)} \tag{5.57b}$$

Hence

$$\left(\frac{1+v}{1-v}\right) \cdot \left(\frac{1+w}{1-w}\right)^2 > \frac{m}{(m-1)} - 2s\frac{n}{(m-1)} \cdot \frac{(1+w)}{(1-v)(1-w)} \cdot \frac{V_T}{V_{in}} \cdot \left(\frac{R_2 + R_1/n}{R_2}\right) \tag{5.58b}$$

If a large value of n/m is required V_{in} should thus be made as large as possible. Usually it is not practicable to exceed the condition $V_{in} = V_T$.

R_2 must be chosen to raise the trigger potential to V_T when m inputs are energized. Thus

$$V_{in} \cdot \frac{m}{n} + \left(\frac{R_1/n}{R_2 + R_1/n}\right) \cdot \left(V_B - V_{in} \cdot \frac{m}{n}\right) = V_T \tag{5.59}$$

Putting $V_{in} \approx V_T \approx \frac{1}{2}V_B$, this leads to

$$\frac{R_1}{n} \approx R_2\left(1 - \frac{m}{n}\right) \tag{5.60}$$

Substituting in Relation (5.58b),

$$\left(\frac{1+v}{1-v}\right)\left(\frac{1+w}{1-w}\right)^2 < \frac{m}{(m-1)}\left[1 - 2s \cdot \left(\frac{2n}{m} - 1\right) \cdot \frac{(1+w)}{(1-v)(1-w)}\right] \tag{5.58c}$$

If the tolerances v and w are reduced to zero, this gives

$$2ms < 1 \Big/ \left(2\frac{n}{m} - 1\right) \tag{5.61}$$

As a first stage in gate design, therefore, a tube must be chosen providing a sufficiently small value of s to satisfy Relation (5.61) with the required values of n and m. Moreover, unless the relation is satisfied with a comfortable margin, the tolerances v and w must be made inconveniently small.

The precise value of R_2 must be chosen to allow for the worst combination of tolerances. Thus

$$(1-v)\,.\,V_{in}\,.\frac{m}{n}\cdot\left(\frac{1-w}{1+w}\right)+\left[V_B-(1-v)\,.\,V_{in}\,.\frac{m}{n}\cdot\left(\frac{1-w}{1+w}\right)\right]\cdot\frac{R_1/n}{R_2+R_1/n}\geqslant(1+s)V_T \quad (5.59a)$$

Until v and w have been determined, however, this Relation cannot be used to obtain R_2. A provisional value for R_2 is therefore obtained first from Equation (5.59). With the aid of Relation (5.58b), values of v and w are chosen so that the limiting value of R_2 may be determined from Relation (5.59a).

Strictly, the tolerances on R_2 and V_B should be considered also. In practice, it is sufficient to ensure that Relation (5.58b) is satisfied by a useful margin. This margin then represents the tolerance on V_B.

Design Procedure for 'm out of n' Gate

(*a*) Set out the following data:
- Total of inputs, n
- Critical number of inputs, m
- Nominal input voltage, V_{in}
- Nominal h.t. voltage, V_B
- Input source impedance, r_{in}
- Fractional tolerance on V_{in}, if known, v

(*b*) Determine the maximum permissible fractional tolerance, $\pm s_{max}$, on the trigger striking potential, V_T.

$$s_{max}=1\Big/\left[2m\left(\frac{2n}{m}-1\right)\right] \quad (5.61a)$$

(*c*) Select a tube for which $s \leqslant 0{\cdot}9s_{max}$.
(*d*) Choose $R_1 \gg r_{in}$.
(*e*) Substitute in Equation (5.59) and solve to obtain an approximate value for R_2.

(*f*) Substitute in Relation (5.58b) and solve for w_{max}.

$$\left(\frac{1+v}{1-v}\right)\cdot\left(\frac{1+w}{1-w}\right)^2<\frac{m}{(m-1)}-2s\cdot\frac{n}{(m-1)}\cdot\frac{(1+w)}{(1-v)(1-w)}\cdot\frac{V_T}{V_{in}}\cdot\left(\frac{R_2+R_1/n}{R_2}\right) \quad (5.58b)$$

(*g*) Substitute in Relation (5.59a) and solve for R_2.
(*h*) Check that the chosen values satisfy Relation (5.57b)
(*j*) If Relation (5.57b) is not satisfied by a margin adequate as a tolerance on V_B, reduce s, v, and/or w until it is.

EXAMPLE 5.7 *'m out of n' Gate*

A gate is required to operate on coincidence of '3 out of 5'. It is proposed that each input be at 100 V, ±5 V, from a 1-kΩ source. Design for 250-V h.t. supply.

(*a*) $n = 5$ $\quad V_{in} = 100$ V
$m = 3$ $\quad v = 0{\cdot}05$
$V_B = 250$ V $\quad r_{in} = 1$ kΩ

(*b*)
$$s_{max}=1\Big/\left[2\times3\times\left(2\times\frac{5}{3}-1\right)\right]=\mathbf{0{\cdot}0716}$$

(*c*) Put $s < 0{\cdot}9s_{max}$
i.e. $s < 0{\cdot}064$

Hence use Z806W or GPE120T ($s \leqslant 0{\cdot}02$)

(*d*) Put $R_1 \gg r_{in}$
i.e. $R_1 = 100$ kΩ

(*e*) Substituting in Equation (5.59), then, for the GPE120T,

$$100\times\frac{3}{5}+\left(\frac{100\div5}{R_2+100\div5}\right)\left(250-100\times\frac{3}{5}\right)=122{\cdot}5$$

$\therefore\ R_2 = 41$ kΩ

(*f*) From Relation (5.58b),

$$\frac{1{\cdot}05}{0{\cdot}95}\times\left(\frac{1+w}{1-w}\right)^2<\frac{3}{2}-2\times0{\cdot}02\times\frac{5}{2}\times\frac{1}{0{\cdot}95}\left(\frac{1+w}{1-w}\right)\times\frac{122{\cdot}5}{100}\times\left(\frac{41+100\div5}{100\div5}\right)$$

$\therefore\ w_{max} = 0{\cdot}034$

Hence, put $w = 0{\cdot}03$

(*g*) Substitute in Relation (5.59a) and solve for R_2.

$$0{\cdot}95 \times 100 \times \frac{3}{5} \times \frac{0{\cdot}97}{1{\cdot}03} + \left[250 - 0{\cdot}95 \times 100 \times \frac{3}{5} \times \frac{0{\cdot}97}{1{\cdot}03}\right] \times \left[\frac{100 \div 5}{R_2 + 100 \div 5}\right] \geqslant 1{\cdot}02 \times 122{\cdot}5$$

$\therefore\ R_2 \leqslant 35{\cdot}0\ \mathrm{k}\Omega$

(*h*) Check that the chosen values satisfy Relation (5.57b).

$$\text{L.H.S.} = 100 \times 1{\cdot}05 \times \frac{2}{5} \times \frac{1{\cdot}03}{0{\cdot}97} + 2 \times 0{\cdot}02 \times 122{\cdot}5 \times \left(\frac{35 + 100 \div 5}{100 \div 5}\right)$$

$$= \mathbf{52{\cdot}3}$$

$$\text{R.H.S.} = 100 \times 0{\cdot}95 \times \frac{3}{5} \times \frac{0{\cdot}97}{1{\cdot}03} = \mathbf{53{\cdot}6}$$

(*j*) Relation (5.57b) allows a tolerance on V_B of only $\pm(53{\cdot}6 - 52{\cdot}3) = \pm 1{\cdot}3V$

Hence reduce w to 0·02

(*h'*) Checking Relation (5.57b) with $w = 0{\cdot}02$,
L.H.S. = **51·2**, R.H.S. = **54·7**

(*j'*) Tolerance on $V_B = \pm(54{\cdot}7 - 51{\cdot}2) = \pm 3{\cdot}5$, say $\pm 3V$.

Solution:

$R_1 = 100\ \mathrm{k}\Omega \pm 2\%$
$R_2 = 35\ \mathrm{k}\Omega \pm 2\%$
$V_B = 250V \pm 3V$
V_1 = GPE.120T

An 'OR' gate may be obtained using the circuits of Fig. 5.20 (*a*) or (*b*) if the input signals separately exceed V_T. Fig. 5.22 (*a*) represents an arrangement which is preferable because the inputs are better isolated from each other and from the output. The diode gate shown in Fig. 5.22 (*b*) offers the same advantages. It is closely related to the 'AND' gate of Fig. 5.20 (*c*) and, in fact, the same design formulae are directly applicable.

By reversing the polarity of one input, the 'AND' gates of Figs. 5.20 (*a*) and (*b*) may be changed to the 'NOT' gates of Figs. 5.22 (*c*) and (*d*). In these gates, the triggering inputs must exceed V_T. The inhibiting inputs should be of amplitude $0{\cdot}5V_T$ to $0{\cdot}7V_T$.

Young [15] has given an account of further logical circuit techniques which represent an extension of the 'ACCESS' pulsed-anode system

referred to on p. 84. Young describes a 'NOT' gate and a single-tube digit store. He discusses also the possibility of a 'NOR' gate. As his circuits rely on the passage of negative trigger current, however, they cannot be recommended for use with high-stability tubes.

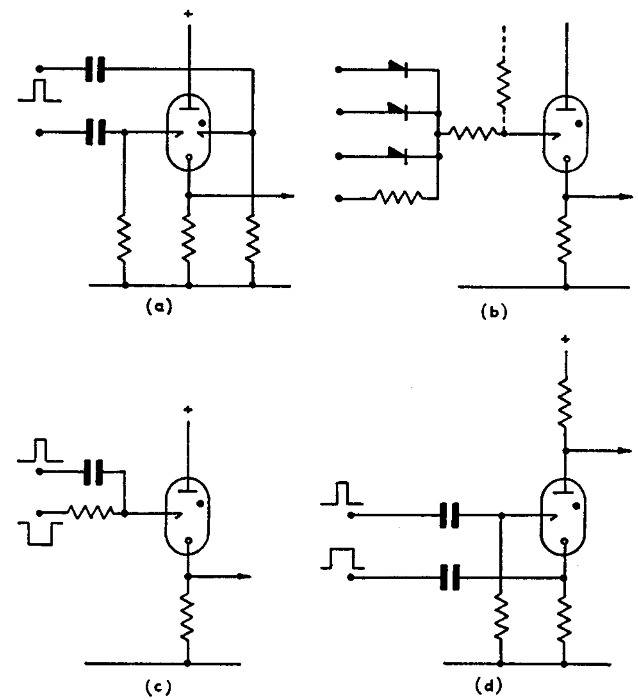

Fig. 5.22. Trigger tube gates: (*a*) Twin-trigger 'OR' gate, (*b*) diode logic (voltage transfer) 'OR' gate, (*c*) bias-before-pulse 'NOT' gate, (*d*) pulse-plus-pulse 'NOT' gate.

Chain and Ring Counters

Fig. 5.23 is representative of a class of counting circuit based on pulse-plus-bias logic. A chain of *n* similar stages is often formed into a closed ring by connecting V_n to V_1 in the same way as V_1 is connected to V_2 and V_2 to V_3. Common-anode-load extinguishing (p. 79) ensures that, in a well-designed circuit, only one tube conducts at a time.

Suppose V_1 is conducting. The voltage drop across R_{K1} raises the trigger of V_2 to a little below triggering potential. The triggers of all other tubes are at earth potential. When a positive-going pulse V_P is applied simultaneously to all triggers it follows that V_2 will fire but, provided $V_P < V_T$, none of the other tubes will fire. Firing of V_2

quenches V_1 and applies a priming bias to V_3. Thus at the next pulse V_3 will strike and V_2 extinguish.

Positive-going outputs may be taken from the cathodes of individual tubes in the ring or chain. Circuits of this type therefore lend themselves to the production of scalers, batching counters, and a wide variety of digital control systems. In some of these applications the cold cathode

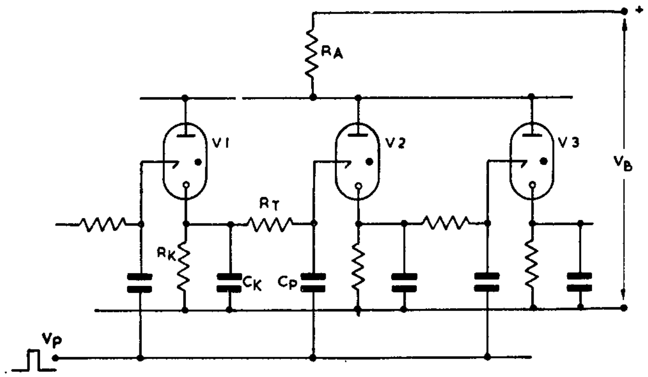

Fig. 5.23. Chain or ring counter using trigger tubes.

stepping tube offers higher speed and greater simplicity. When special logical functions must be performed, however, the chain counter is often a better choice.

In designing a chain counter particular care must be paid to component tolerances. Flood and Warman [10] have demonstrated that tubes with oxide-coated cathodes show such wide tolerances in trigger characteristics that little or nothing can be allowed for other component tolerances. Tubes with pure metal cathodes are to be preferred for their better stability through life. Even these can show considerable drift in trigger striking voltage due to the passage of reverse trigger current. If, for example, the counter is left static with V_2 conducting, then the extinction of V_1 will cause V_2 to lose its priming bias. The trigger acts as a probe in the anode–cathode discharge, and its potential approximates the anode maintaining voltage. Consequently, a negative trigger current flows from the trigger of V_2 to ground, its value being limited by R_T and R_K. This reverse current causes electrode material to be sputtered from the trigger on to the cathode, so raising the value of V_T. It is best to design the stage so that, with the counter static, the tube current substantially fills the cathode – say, $I_K \approx 0{\cdot}8 I_{K(av)\,(max)}$. This ensures that foreign material will continually be sputtered off the cathode.

To particularize, the normal value of V_T for the Z700U is in the range 137–153 V. To allow for the effects of reverse trigger current, however, the design should preferably accommodate an increase of V_T to 175 V. Reliable low-frequency operation therefore requires $V_K < 135$ V and $V_P > 40$ V. At higher stepping rates the full value of V_K is not available as a bias at the adjacent trigger before the arrival of the next pulse. Both V_K and V_P must therefore be made as large as is permissible.

The time-constant $C_P R_T$ further restricts the rate at which trigger biases can rise and fall. Too great a value will too greatly reduce the maximum stepping frequency, f_{max}. Too small a value leads to differen-

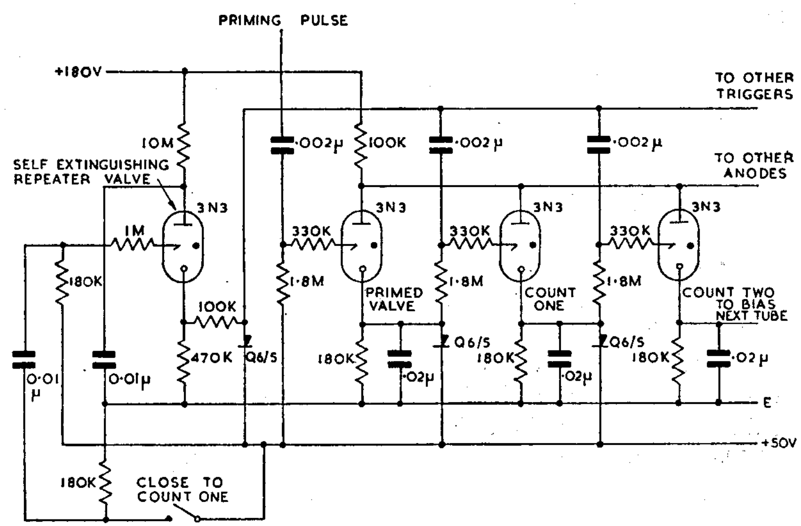

Fig. 5.24. Chain counter using rectifier clamps.

tiation of the pulse V_P so that a negative-going pulse is applied to unbiased tubes. When the tube V_n is conducting and a further pulse is applied to strike $V_{(n+1)}$ there is thus a danger of false triggering of $V_{(n-1)}$ due to the combination of this negative pulse and the residual positive bias still on its cathode. Published designs usually show a compromise value of $C_P R_T \approx T_D$, and it follows then that $f_{max} \approx 1/2T_D$. The value of C_P should correspond to the minimum value of C_T for reliable capacitor triggering. R_T will then be as large as possible and will generally restrict the reverse trigger current to a safe value when the counter is static.

In a ring counter a rigorous design must allow for the 'worst × worst'

combinations of tolerances on V_B, V_P, V_T, V_M, R_A, and R_K, all of which interact. Even with molybdenum-cathode tubes, this is likely to call for 5% tolerances on the values of V_B, R_A, and R_K. If preferred values are to be used it may be necessary to use still closer tolerances. It is, of course, possible to ease the tolerance on one parameter at the expense of a closer tolerance on another.

Counter design may be eased considerably by the use of catching diodes. In Fig. 5.24 rectifier clamps are used to limit the priming bias to the voltage on a reference rail. As a result, the bias is substantially independent of the interval between stepping pulses. The rectifier diodes must withstand about 150 V, however, and the additional cost per stage is significant. Additionally, or alternatively, a single catching diode may be used to limit the amplitude of the drive pulse, V_P. At a relatively small cost this removes one of the variables and so eases the problem of design.

EXAMPLE 5.8 Chain or Ring Counter

Tube manufacturers commonly publish designs for counting circuits using their own trigger tubes. The user will probably prefer to adopt one of these designs rather than perform the rather laborious design procedure himself. Nevertheless, an understanding of design procedure can be valuable, and the following example illustrates how the component values may be arrived at for a Mullard design using the Z700U in the circuit of Fig. 5.23. Values and tolerances are determined for reliable operation over the widest possible range of stepping speeds.

(*a*) Listing the principal data on the Z700U:

(*a*) $V_{IG} = 310$ V	$T_D = 0{\cdot}1$ msec (approx)
$V_T = 145$ V ± 8 V	$T_I = 20$ µsec for $V_0 = 20$ V
$V_M = 116$ V ± 5 V	$I_{K(pk)\,(max)} = 16$ mA
$V_N = 115$ V	$I_{K(av)\,(max)} = 4$ mA

(*b*) Determine V_B. Tentatively putting the fractional tolerance v on V_B as 0·05, then to ensure that no tube strikes before the first tube is triggered,

$$V_B < V_{IG}/(1 + v) = 310/1{\cdot}05 = 295 \text{ V}$$

Hence, provisionally, $V_B = 280 \text{ V} \pm 5\%$

(*c*) Determine R_A.

$$R_A \geqslant (V_{B(max)} - V_{M(min)})/I_{K(pk)\,(max)} \qquad (5.62)$$
$$= (294 - 111)/16 = 11{\cdot}4 \text{ k}\Omega$$

For good tube life the peak current should lie between $I_{K(av)\,(max)}$ and $I_{K(pk)\,(max)}$. Hence, put $R_A = 18$ kΩ.

(*d*) To provide a high maximum stepping speed without false triggering at low speeds,

$$V_{K(max)} \approx V_{T(min)} \qquad (5.63)$$

i.e.
$$\frac{R_K}{R_A + R_K} \cdot (V_{B(max)} - V_{M(min)}) \approx V_T$$

$$\therefore \frac{R_A}{R_K} = \frac{(V_{B(max)} - V_{M(min)})}{V_{T(min)}} - 1 \qquad (5.64)$$

$$= \frac{294 - 111}{137} - 1 = 0{\cdot}336$$

$$\therefore R_K = R_A/0{\cdot}336 = 18/0{\cdot}336 = 53{\cdot}6 \text{ k}\Omega$$

To put $R_K = 47$ kΩ would reduce V_K and so restrict the maximum stepping speed. Accordingly, the design is continued with $R_K = 56$ kΩ on the supposition that this can be accommodated by adopting a closer tolerance on V_B.

(*e*) For a chosen fractional tolerance, $w = 0{\cdot}10$, on the resistors R_A and R_K, determine the permissible fractional tolerance, v, on V_B.

$$V_{K(max)} = [(1 + v)V_B - V_{M(min)}] \cdot \frac{(1 + w)R_K}{(1 + w)R_K + (1 - w)R_A} \qquad (5.65)$$

$$\therefore v \leqslant \left\{\frac{(1 + w)R_K + (1 - w)R_A}{(1 + w)R_K} \cdot \frac{V_{T(min)}}{V_B} + \frac{V_{M(min)}}{V_B}\right\} - 1 \qquad (5.66)$$

$$= \left\{\frac{1{\cdot}10 \times 56 + 0{\cdot}90 \times 18}{1{\cdot}10 \times 56} \times \frac{137}{280} + \frac{111}{280}\right\} - 1 = 0{\cdot}013$$

Hence put $v = 0{\cdot}0125$, i.e. $V_B = 280 \text{ V} \pm 1{\cdot}25\%$.

(*f*) Determine the value of C_K.

$$R_K/R_A = 56/18 = 3{\cdot}1$$

From Fig. 4.14, $\alpha T = 0{\cdot}33$

$$\therefore C_{K(min)} = \frac{T_D}{\alpha T \,.\, R_K} = \frac{0{\cdot}1 \times 10^{-3}}{0{\cdot}33 \times 56 \times 10^3} = 5{,}400 \text{ pF}$$

$\therefore$ put $C_K =$ **6,800 pF ±20%**

(*g*) With the counter static,

$$V_{A(min)} = V_{B(min)} - \frac{(1 + w)R_A}{(1 + w)R_A + (1 - w)R_K} \cdot [V_{B(min)} - V_{M(max)}] \qquad (5.67)$$

$$= 276{\cdot}5 - \frac{1{\cdot}1 \times 18}{1{\cdot}1 \times 18 + 0{\cdot}9 \times 56} \times (276{\cdot}5 - 121) = \mathbf{232{\cdot}6\ V}$$

(*h*) Determine C_P.

$$C_{T(crit)} \approx \frac{I_T T_I}{V_T - V_N} \qquad (4.6)$$

When $V_A = 232$ V, $I_{T(max)} \approx 40$ μA,

$$\therefore\ C_{T(crit)} \approx \frac{40 \times 10^{-6} \times 20 \times 10^{-6}}{145 - 115} = 26{\cdot}7 \text{ pF}$$

Putting $C_P \geqslant 3C_{T crit)}$, $C_P = \mathbf{100\ pF} \pm 10\%$

(*j*) Put $C_P R_T \approx T_D = 0{\cdot}1 \times 10^{-3}$ sec.

Thus $R_T \geqslant 1$ MΩ. To allow for tolerance on C_P, put $R_T = \mathbf{1{\cdot}2\ M\Omega} \pm \mathbf{10\%}$

(*k*) Consider the minimum interval, τ, between pulses consistent with reliable operation:

$$V_{C(crit)} + V_P = V_T + V_0 \tag{4.8}$$

where $V_{C(crit)}$ is the bias available on the trigger at the time τ after striking the preceding tube.

Put
$$C_K \cdot \frac{R_A R_K}{R_A + R_K} = T_1$$
and
$$C_P R_T = T_2$$
Then, if $R_T \gg R_K$, and $T_1 \neq T_2$,

$$V_{C(crit)} \approx V_{K(pk)} \left\{1 - \frac{T_1}{T_1 - T_2} \cdot e^{-\tau/T_1} + \frac{T_2}{T_1 - T_2} \cdot e^{-\tau/T_2}\right\} \tag{5.68}$$

Thus,

$$V_P + V_{K(pk)} \left\{1 - \frac{T_1}{T_1 - T_2} \cdot e^{-\tau/T_1} + \frac{T_2}{T_1 - T_2} \cdot e^{-\tau/T_2}\right\} = V_T + V_0 \tag{5.69}$$

When the counter has been static and a train of pulses is then applied separated by time τ, the second pulse will cause spurious retriggering of the formerly conducting tube if the value of V_P is excessive. Due to differentiation of the input pulse, this tube receives a negative-going trigger pulse given by:

$$V_C' = V_P \,.\, e^{-T_P/T_2} \tag{5.70}$$

where T_P = pulse duration.

At the instant in question, the cathode capacitor has been discharging through R_K for a time $(\tau + T_P)$. Putting $C_K R_K = T_3$, the cathode potential is given by:

$$V_K' = V_{K(pk)} \,.\, e^{-(\tau + T_D)/T_3} \tag{5.71}$$

In the limiting case,

$$V_C' + V_K' = V_T + V_0$$

$$\therefore\ V_P \,.\, e^{-T_P/T_2} + V_{K(pk)} \,.\, e^{-(\tau + T_D)/T_3} = V_T + V_0 \tag{5.72}$$

Thus, if one puts $T_P = T_I$, Equations (5.69) and (5.72) provide two simultaneous equations with two unknowns: V_P and τ. They may thus be combined and solved graphically to give $V_P \approx 100$ V and $\tau \approx 0{\cdot}2$ msec, whence $f_{max} \approx 5$ kc/s.

Solution:

$V_B = 280$ V $\pm 1{\cdot}25\%$	$C_K = 6{,}800$ pF $\pm 20\%$
$V_P = 100$ V $\pm 5\%$	$C_P = 100$ pF $\pm 10\%$
$R_A = 10$ kΩ $\pm 10\%$	$T_P = 20$ μsec
$R_K = 56$ kΩ $\pm 10\%$	$f_{max} = 5$ kc/s
$R_T = 1{\cdot}2$ MΩ $\pm 10\%$	

If the Z700U of V_0 is replaced by a Z700W the additional trigger of this tube may be used for resetting the counter to zero.

Reversible Counters

By using double-trigger tubes for all stages, a reversible counter may be constructed. One trigger of each tube is capacitor-coupled to a source of 'add' pulses. Each 'add' trigger is biased from the cathode of the preceding tube, as shown in Fig. 5.23. Each 'subtract' trigger is similarly biased from the cathode of the succeeding tube. 'Add' and 'subtract' pulses cannot normally be applied concurrently, as there is a danger of false triggering if pulses are separated by a time less than $1/f_{max}$. The same problem arises with the forced resetting mentioned above. Neglect of this restriction can lead to the simultaneous conduction of two tubes in a single ring.

Modified Chain Counters

Sidorowicz [11, 12] has described several variants of the chain or ring counter designed to meet particular requirements. They include a staircase waveform generator, a pulse distributor, and a reversible counter. He describes also alternative forms of decade ring counter. One of these uses fourteen tubes to provide increased operating speed, others effect an economy of components by using only five or six tubes instead of the usual ten. Considerable ingenuity has been shown in devising arrangements which use to best advantage tubes having oxide-coated cathodes. In view of the comments of Flood and Warman [10] on the problems of component tolerancing with such tubes, there is little doubt that most of these circuits could be made to perform with greater speed and reliability if redesigned for tubes using pure metal cathodes.

In this connexion it may be relevant that Sidorowicz indicates difficulties arising from ionic coupling between triggers in a reversible ring counter using twin-trigger tubes. Apparently no such problems have been met by Liebendorfer [3] and others who have published similar circuits for twin-trigger tubes using pure metal cathodes.

Rings of ten tubes may be cascaded to provide decade scalers. It is then necessary to provide special means for transferring a 'carry' pulse from one ring to the next. This may be done by a pulse amplifier comprising a self-quenching tube capable of supplying a low-impedance pulse of requisite amplitude to the triggers of the succeeding decade. Usually this tube is triggered in parallel with tube '0' of the units ring. Sidorowicz [12] shows an unusual arrangement in which the pulse amplifier is placed between tubes '9' and '0'. It may be argued that if the pulse amplifier fails to operate on all occasions the errors will be numerically smaller with this arrangement.

Batching Counters

A range of cold cathode equipment produced by Mullard uses pulse-plus-bias logic circuits to indicate when decade ring counters have

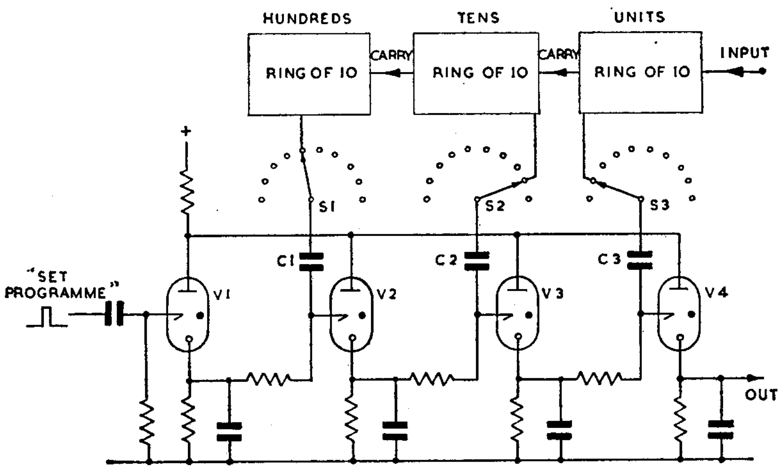

Fig. 5.25. Programme tubes, V_1–V_4, indicating completion of pre-selected count in decade ring counter at top of diagram.

reached one or more particular pre-selected counts. It will be seen from Fig. 5.25 that the 'programme' tubes, V_1 – V_4, are used in arrangement very similar to that of a chain counter. Switches S_1, S_2, and S_3 connect the capacitors to selected cathodes of the tubes in the ring counters. Thus, supposing a count of '481' is to be detected, switches S_1, S_2, and S_3 are set to connect their capacitors to the cathodes of tubes '4', '8', and '1' in their respective decade counters.

A programme is initiated by applying a pulse to fire V_1. As tubes

V_1–V_4 share a common anode load, firing of V_1 extinguishes any other conducting tube. At the same time the rise in cathode potential of V_1 causes a bias to appear on the trigger of V_2. If now a series of pulses is applied to the decade scaler, C_3 will apply a positive pulse to the trigger of V_4 each time tube '1' of the 'units' decade is struck. In the absence of a bias from V_3, however, the pulse is of insufficient amplitude to fire V_4. Similarly, pulses are applied by C_2 to V_3 each time tube '8' is struck in the 'tens' decade. Since there is no bias on V_3, however, these pulses do not produce triggering. Thus no further programme tubes strike until the decade scaler reaches a count of '400'. As tube '4' of the 'hundreds' decade strikes, its cathode delivers a positive pulse via S_1 and C_1 to the trigger of V_2. Since V_1 has already applied a bias to V_2, V_2 fires and, in so doing, quenches V_1 and applies a bias to V_3. When the count reaches '480' a pulse passes via S_2 and C_2 to strike V_3. V_3 now quenches V_2 and biases V_4. Finally, at a count of '481', V_4 is fired by a pulse received via S_3 and C_3. Thus an output appears on the cathode of V_4 as soon as the pre-selected count has been reached.

Many industrial applications may be found for this type of circuit. It may be used to count articles into boxes by batches of one hundred, one gross, or any other required number. Each time an output appears on V_4 the flow of articles is diverted to the next box and the counter is reset to zero.

By duplicating the switches S_1, S_2, and S_3 and the programme tubes V_1–V_4, a digital speed control system may be constructed. A tachometer probe applies a pulse input to the decade counter, and after a pre-determined time interval the state of the programme outputs indicates whether the speed is low, within limits, or high. If the speed is within the pre-selected limits the counter will have passed one of the pre-selected numbers but not the other. With still more programme channels, speed errors may be classified according to degree as well as kind.

Pattern Shifting Registers

Fig. 5.26 shows a pattern shifting register which bears an obvious similarity to the chain counter of Fig. 5.23. R_B has been added between R_K and C_K, and a negative-going pulse is applied to the h.t. rail just before the positive-going pulse is applied to the triggers. By these means, a circuit is obtained in which any combination of tubes may be conducting and in which the pattern of conducting tubes is displaced one position to the right whenever the h.t. rail and triggers are pulsed.

Like the chain counter, this circuit relies on pulse-plus-bias logic. Suppose only V_2 is conducting. The cathode of V_2 will rise to a potential

$V_K = (V_B - V_M)$. This bias potential is applied to the trigger of V_3. The negative-going pulse on the h.t. rail is long enough to allow V_2 to deionize, but so short that much of the bias remains on V_3 when the trigger is pulsed. Consequently, when the two pulses are applied V_3 is triggered and the pattern of conduction is displaced to the right. By making $R_B \gg R_K$, it is ensured that as soon as any tube is extinguished C_K applies to the cathode only a small proportion of the positive bias V_K which previously appeared across R_K. Thus, whereas in a chain counter C_K prevents a tube from restriking, in the shift register this does not occur. If, in the above example, V_1 had been conducting as well as V_2, then after one transfer pulse V_2 would have restruck due to the combination of trigger pulse and the bias previously received from V_1. Thus the initial pattern, 'V_1 on, V_2 on, V_3 off' would become 'V_1 off, V_2 on, V_3 on'.

Design of a shift register is rather more complex than that of a chain counter, since additional variables arise in the width of the anode quenching pulse and the time interval between it and the trigger pulse. No formal design procedure is offered here, in fact, since this must depend on which of several alternatives is adopted. For example, the trigger pulse generated by V_Z in Fig. 5.26 will vary in amplitude with variations in V_B. A design using pulses of constant amplitude would be more straightforward and would lead to a larger tolerance on the value of V_B.

The circuit of Fig. 5.26 was, in fact, designed to use a change-over microswitch to provide the h.t. interruption. As the switch arm moves from one contact to the other, the h.t. is interrupted long enough to allow the conducting tube to deionize. On reaching the opposite contact, however, the moving contact bounces for several milliseconds. It is thus necessary to delay the generation of the trigger pulse until the bouncing of the microswitch contact has ceased. This delay, which is provided by R and C in the trigger circuit of V_Z, leads to a considerable restriction of the circuit performance. As a result, the design shown has a maximum stepping rate of only about 2 shifts/sec and can accept an h.t. supply variation of no more than ± 10 V on a nominal 225 V.

As a guide to design, the following principles may be indicated:

The maximum bias which may be applied to a tube must always be less than V_T. Thus,

$$V_K = V_{B(max)} - V_{M(min)} < V_{T(min)} \tag{5.73}$$

Whence,

$$V_{B(max)} < V_{T(min)} + V_{M(min)} \tag{5.73a}$$

Similarly, the maximum trigger pulse amplitude, V_{PT}, must not exceed $V_{T(min)}$, but must approach V_T if performance is not to suffer. Close-tolerance tubes are therefore desirable.

The arrangement of V_Z shown in Fig. 5.26 would appear to provide a pulse of amplitude $(V_B - V_M)$. In practice, the pulse is rounded by

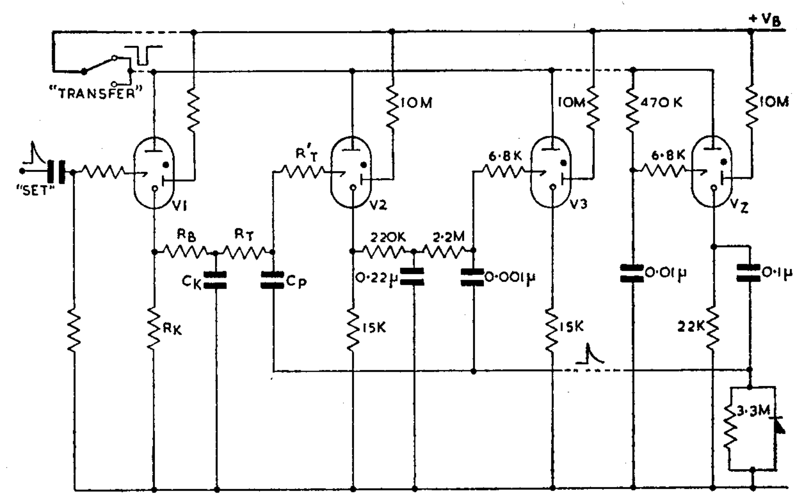

Fig. 5.26. Pattern shifting register.

the finite formative delay of V_Z, and so some 15–20 V of amplitude is lost in the coupling capacitors.

The time-constant $C_P R_T$ must be small enough to provide extinction of the trigger–cathode gap during the anode quenching period. Thus

$$C_P R_T \leqslant T_D$$

where T_D = deionization time.

$$(\text{In Fig. 5.26, } C_P R_T \approx \tfrac{1}{2} . T_D)$$

An increase in delay time, τ, between the leading edges of the anode and trigger pulses requires an increase in $C_K R_B$ if $V_{B(min)}$ is not to be increased. On the other hand, rapid cycling demands that $C_K R_B$ be kept sufficiently low to ensure that the bias required for one cycle shall have collapsed almost entirely before the next.

If values of the maximum required rate of cycling, $f_{c(max)}$, and τ are known, a first estimate of $C_K R_B$ may be made from Equation (5.74).

$$(C_K R_B)_{optimum} \approx \left(\frac{\tau}{f_{c(max)}}\right)^{\frac{1}{2}} \tag{5.74}$$

With this value, the maximum tolerance on V_B should be obtained.

Hough and Ridler [13] have described an alternative form of shift register.

Self-resetting Circuits

Crowther [2, 14] has described a number of self-resetting circuits useful where a relay must operate when an input exceeds a critical level and release when the input falls below the same level. Such circuits use either

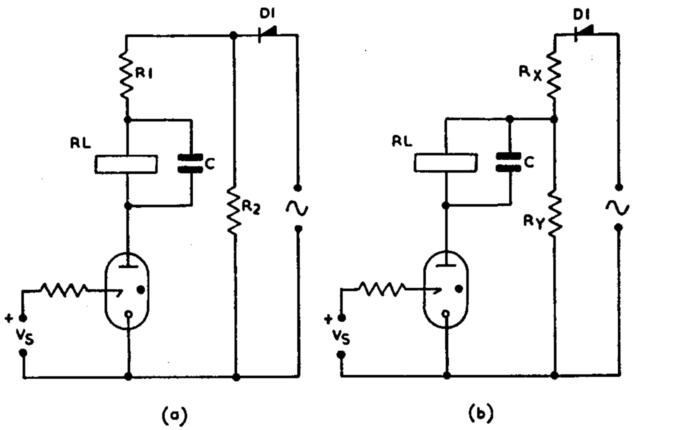

Fig. 5.27. Self-resetting relay circuits (*a*) for $V_m\sqrt{2} < V_{IG}$, (*b*) for $V_m\sqrt{2} > V_{IG}$.

a.c. or unsmoothed rectified a.c. to supply the anode. Provided the trigger potential is sufficiently positive, the tube strikes each time the anode potential rises, and subsequently extinguishes as the anode falls below the maintaining potential. The pulsations of the intermittent anode current may be smoothed by a capacitor connected across the load.

The simple circuit shown in Fig. 5.27 shows a backlash between the values of input causing the relay to operate and release. If a d.c. input V_S is applied to the trigger and gradually increased the tube will fire on every positive half-cycle of the anode supply when $V_S > V_T$. The trigger–cathode discharge is not quenched when the anode potential falls, however. Consequently, V_S must be reduced below the trigger-maintaining voltage, V_N, before the tube will cease to restrike on each new cycle. The backlash in critical input voltage is thus (($V_T - V_N$).

Backlash may be reduced by quenching the trigger discharge before each new cycle so that only V_T is of importance. This may be done by

making the trigger circuit self-quenching. Such an arrangement might appear to require only a trigger–cathode capacitor in Fig. 5.27. A difficulty arises, however, in that the trigger discharges may either fail to synchronize with the anode waveform or else may synchronize unfavourably. In the first event the mean anode current will fluctuate undesirably. In the second case the trigger may discharge just before the anode voltage has risen sufficiently for anode–cathode breakdown to occur. The next trigger discharge may then occur after the anode potential has again fallen so that no anode–cathode discharge occurs at all, even though the trigger is discharging twice in every supply cycle.

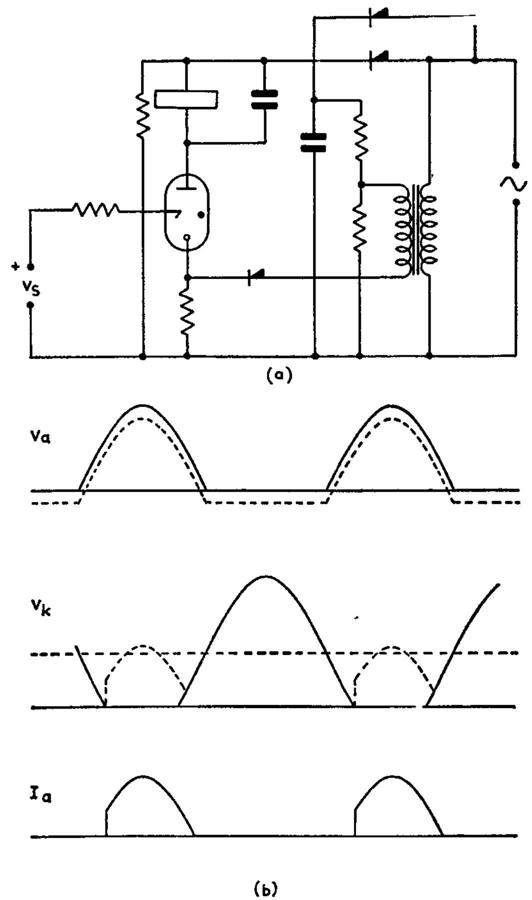

Fig. 5.28. Self-resetting relay with trigger discharge synchronized and quenched by a.c. to reduce backlash.

Crowther [2] has discussed this effect in detail and has shown that one solution lies in returning the trigger capacitor not to ground, but to a tapping on the anode supply. The trigger discharge is thereby synchronized to occur in the appropriate phase. Variations in anode supply voltage will affect the critical input voltage, however, and this technique is therefore not applicable where the greatest precision is required.

An alternative method of synchronizing and quenching the trigger discharge is shown in Fig. 5.28. A combination of d.c. and a.c. bias is applied to the cathode so that when the anode discharge extinguishes, the cathode is at the same time carried positive to extinguish the trigger

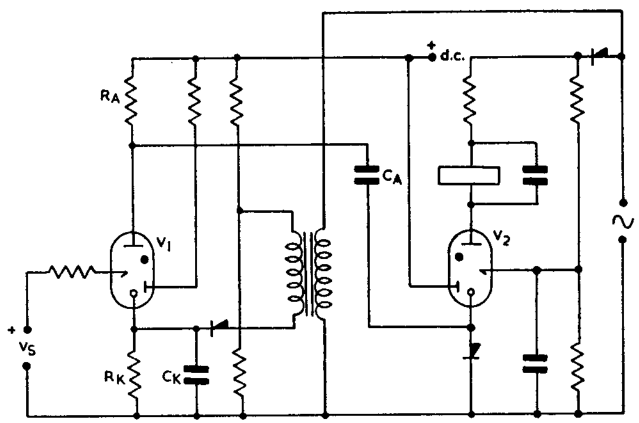

Fig. 5.29. Self-resetting relay in which dissipation in V_1 is kept low to minimize trigger hysteresis effect.

discharge. On each new cycle, however, the diode becomes non-conducting just after the anode has reached working potential. While the diode is non-conducting there is no bias on the cathode, and a trigger discharge will occur if the trigger potential is at or above V_T. When the tube fires, cathode current flows through the bias resistor to produce positive-going excursions as shown dotted in the waveforms of Fig. 5.28 (*b*).

Even the circuit of Fig. 28 shows a backlash of 1–5 V. This is due to the hysteresis effect (p. 67) arising from internal heating of the tube when current is drawn. This may be minimized by using a tube designed to show only a small hysteresis effect and/or by arranging that the tube dissipation is as small as possible.

In Fig. 5.29, V_1 operates in the self-quenching mode and hence carries

a mean anode current of $<100\ \mu$A. Since this is too small a current to operate a robust relay directly, V_1 is made to trigger a second tube, V_2, by applying a negative-going pulse to its cathode (p. 75). The previously described cathode biasing system is shown synchronizing the discharge of V_1 to the anode supply of V_2. The insertion of R_K in the path of the self-extinguishing discharge has no deleterious effect provided it is by-passed by C_K and $C_K \gg C_A$. It is desirable also to put $C_K R_K \ll 1/f$, where f is the a.c. supply frequency.

Crowther and Gimson [14] have described a simplified version of this circuit, suitable for a.c. inputs. With two of these circuits they have produced an automatic a.c. voltage controller of the motor-driven variable-transformer type.

For a wide range of applications it is not necessary to go to these lengths. The simple current-triggering arrangement shown in Fig. 5.27 is adequate for those touch-circuits which provide a large change in input level between the 'on' and 'off' conditions.

Even when discrimination is required between input levels only slightly above and below V_T, this may be provided by designing the trigger circuit so that self-quenching discharges are produced at a sufficiently high frequency.

Hence

$$C_T R_T \ll 1/f \qquad (5.75)$$

The value of C_T must exceed the minimum specified by the manufacturer, $C_{T(min)}$.

$$\therefore\ r_{in} + R_T \leqslant C_T/100f \qquad (5.76)$$

where r_{in} = input source impedance,
f = frequency of a.c. mains supply.

To prevent the anode potential from becoming excessively negative during the non-conducting half-cycle, it is in many cases sufficient to provide a rectifier load (R_2 in Fig. 5.27 (*a*)) of resistance much lower than the diode back-resistance. When shielded-anode tubes are used (Fig. 5.30) the shield supply potentiometer, R_3 and R_4, serves this purpose. With high-resistance relays the drop across the relay, $I_{K(mean)} R_L$, can exceed the permissible negative voltage which may be applied to the anode. Fig. 5.30 (*b*) shows one solution to this problem. During the negative half-cycle of the supply D_2 becomes non-conducting. The voltage across R_L is then divided between a negative voltage at the tube anode and a positive voltage on R_1, proportioned in the ratio $R_6 : R_5$. Atkinson [16] recommends D_2 = OA202, $R_5 = 1\ \text{M}\Omega$, $R_6 = 470\ \text{k}\Omega$.

In Figs. 5.27 and 5.30 the priming anodes have been omitted for simplicity of presentation. Ideally they should be returned through the appropriate resistances to a positive d.c. supply. If they are returned to

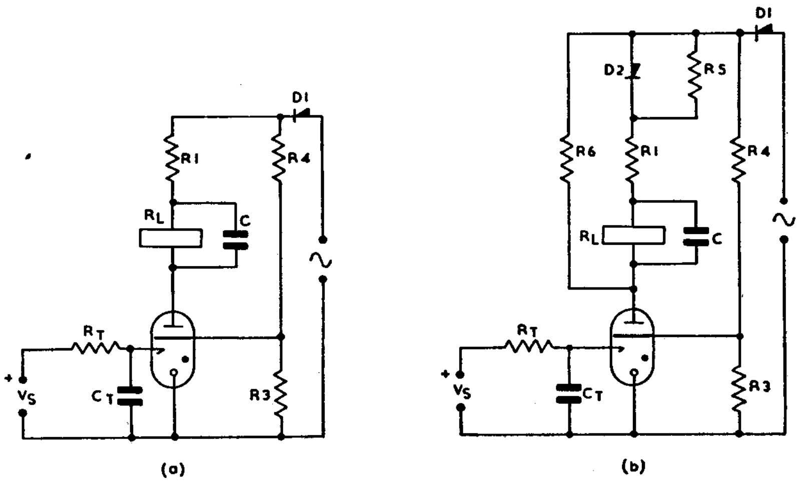

Fig. 5.30. Self-resetting relays using shielded-anode tubes. (*a*) Basic circuit, (*b*) with additional components required with high voltage drop across relay R_L.

the half-wave rectified supply there may be some jitter in the firing angle of the main anode due to statistical delays in striking of the priming anode discharge.

Design of Anode Circuit

Atkinson [16] has published a set of curves from which the optimum anode circuit for the Z806W may be determined. He takes as his worst case the condition of low supply voltage, high resistor tolerance, and anode conduction delayed until the peak of the supply voltage waveform. This last eventuality halves the mean current drawn by the tube.

In the general design procedure given below it is assumed that anode conduction occurs during the whole of that part of the cycle over which the supply voltage exceeds the sum of the anode maintaining voltage, V_M, and the drop, $I_{K(\text{mean})}R_L$, across the relay. Provided the anode take-over voltage (i.e. the minimum anode voltage at which transfer to the anode–cathode gap will occur) is less than 80% of the peak supply voltage, the errors introduced by this assumption may be covered by designing for a relay pull-in current 10% higher than the true value. The only assumption which remains is that a trigger discharge will be

present at the moment the anode voltage reaches the anode take-over potential, i.e. (*a*) if current triggering is used the trigger current must exceed the transfer current at the anode take-over potential, or (*b*) if capacitor triggering is used the discharge is synchronized to occur in the optimum phase, or (*c*) if capacitor triggering is used, self-quenching discharges occur at a frequency at least one hundred times greater than the supply frequency.

In the extreme condition considered by Atkinson a significant delay occurs between the application of an anode voltage exceeding the value for take-over and the arrival of the next trigger discharge. The general design procedure may be made to cover this most pessimistic case by designing for a relay pull-in current double the true value.

Estimation of the mean and peak cathode currents and a suitable value for R_1 resembles the design of a trigger tube stabilizer (pp. 102 *et seq*). In this case, however,

$$k = V_m\sqrt{2}/(V_m + g \,.\, I_{\text{rel}}R_L) \tag{5.77}$$

and

$$p = u \,.\, I_{K(\text{peak})}/g \,.\, I_{\text{rel}} \tag{5.14a}$$

where V_m = r.m.s. value of the a.c. supply,
$u = \frac{1}{2}$ if half-wave rectification is used, or
$u = 1$ if full-wave rectification is used,
I_{rel} = relay pull-in current,
g = factor (between 1 and 2) by which I_{rel} must be increased to allow for considerations discussed above.

Now,

$$I_{K(\text{peak})} = \frac{1}{R_1}\left\{V_m\sqrt{2} - (V_m + g \,.\, I_{\text{rel}}R_L)\right\} \tag{5.78}$$

Substituting from Equations (5.77) and (5.14a) for $(V_M + g \,.\, I_{\text{rel}}R_L)$ and $I_{K(\text{peak})}$ respectively,

$$R_1 = \frac{u(k-1)}{k \,.\, p} \cdot \frac{V_m\sqrt{2}}{g \,.\, I_{\text{rel}}} \tag{5.79}$$

V_m should be chosen to be as high as possible without exceeding V_{IG} at the crest of the waveform. g is a function of the ratio between anode take-over voltage (V_{AT}) and V_m. The table on p. 153 allows the approximate value of g to be determined. Hence k may be determined from Equation (5.77) and p is thereafter obtained from Fig. 5.7.

For trigger tubes not provided with a shield anode, supply voltages in excess of about 200 V cannot normally be used without resource to the circuit of Fig. 5.27 (*b*) in order to restrict the effective value of V_m to a lower value, V_m'. R_1 is then provided by the impedance of R_X and R_Y in parallel.

In adjusting R_X and R_Y to preferred values, care must be taken neither to increase V_m' (so that anode–cathode breakdown occurs) nor to reduce V_m' so far that the required value of I_K is not provided in all circumstances. Equation (5.79) may be used to calculate the maximum permissible value of R_1. Some tolerance on V_m' may be obtained by then choosing R_X and R_Y to give a lower value of R_1.

From Equations (5.77) and (5.79) it may be deduced that, if R_1 and V_m' are changed in a manner which keeps $g \,.\, I_{rel}$ constant,

$$\frac{dR_1}{R_1} = \frac{k}{k-1} \cdot \frac{dV_m'}{V_m'} \tag{5.80}$$

In a typical circuit $k \gtrsim 2{\cdot}0$, and hence $k/(k-1) \lesssim 2{\cdot}0$, i.e. a 10% reduction in R_1 allows 5% reduction in V_m'.

When it is necessary to use the circuit of Fig. 5.27 (*b*) it must be remembered that the tolerance, t, on R_X and R_Y affects the ratio $R_Y/(R_X + R_Y)$. Putting

$$\rho = \frac{V_m'}{V_m} = \frac{R_Y}{(R_X + R_Y)} \tag{5.81}$$

$$\frac{1}{\rho_{min}} = \left(\frac{V_m}{V_m'}\right)_{(max)} = 1 + \frac{R_X}{R_Y} \cdot \frac{(1+t)}{(1-t)} \tag{5.81a}$$

and

$$\frac{1}{\rho_{max}} = \left(\frac{V_m}{V_m'}\right)_{(min)} = 1 + \frac{R_X}{R_Y} \cdot \frac{(1-t)}{(1+t)} \tag{5.81b}$$

$$\therefore \frac{\dfrac{1}{\rho_{min}} - \dfrac{1}{\rho_{max}}}{1/\rho} = \frac{\left(\dfrac{1+t}{1-t}\right) - \left(\dfrac{1-t}{1+t}\right)}{(R_X + R_Y)/R_Y} \cdot \frac{R_X}{R_Y}$$

$$= \frac{4t}{1-t^2} \cdot \frac{R_X}{(R_X + R_Y)} \tag{5.82}$$

Now

$$\frac{d(1/a)}{1(/a)} = \frac{-da}{a}$$

$$\therefore \frac{\delta\rho}{\rho} = \frac{\rho_{max} - \rho_{min}}{\rho} = \frac{4t}{1-t^2} \cdot \frac{R_X}{R_X + R_Y} \tag{5.82a}$$

But

$$\frac{dV_m'}{V_m'} = \frac{d\rho}{\rho}$$

Hence, combining Equations (5.80) and (5.82a),

$$\frac{\delta R_1}{R_1} = \frac{k}{k-1} \cdot \frac{4t}{1-t^2} \cdot (1-\rho) \tag{5.83}$$

In general, $t \ll 1$ and the tolerance, t, on R_X and R_Y is given by

$$t \simeq \frac{1}{4} \cdot \frac{k-1}{k} \cdot \frac{1}{1-\rho} \cdot \frac{\delta R_1}{R_1} \tag{5.83a}$$

For a shielded-anode tube the manufacturer usually specifies the appropriate proportion, z, of the anode supply voltage to be applied to the shield, A_2. Thus, referring to Fig. 5.30,

$$z = R_3/(R_3 + R_4) \tag{5.84}$$

and

$$R_{A2} = R_3R_4/(R_3 + R_4) \tag{5.85}$$

Whence,

$$R_4 = R_{A2}/z \tag{5.86}$$

and

$$R_3/R_4 = z/(1-z) \tag{5.87}$$

Design Procedure for Self-resetting Relay

(*a*) Set out the following data for the selected tube:
- Anode breakdown potential, V_{IG}
- Anode maintaining potential, V_M
- Maximum permissible average cathode current, $I_{K(av)\,(max)}$
- Maximum permissible peak cathode current, $I_{K(pk)\,(max)}$
- Minimum anode voltage for transfer (anode take-over voltage), V_{AT}
- Nominal impedance of supply to shield anode, R_{A2}

(*b*) Set out further design data:
- R.m.s. voltage of a.c. supply, V_m
- Fractional tolerances, $+r_1$ and $-r_2$, on a.c. supply voltage
- Fractional tolerance $\pm w$ on R_1 (Figs. 5.27 (*a*) and 5.30)
- Supply frequency, f
- Full-wave rectification ($u = 1$), or half-wave ($u = \frac{1}{2}$)

(*c*) Determine r.m.s. value, V_m', of required nominal a.c. supply voltage,

$$V_m' \leqslant \frac{V_{IG}}{(1 + r_1)\sqrt{2}} \tag{5.88}$$

(*d*) Determine the minimum factor, g, by which the relay pull-in current must be increased,

V_{AT}/V_m'	g
0·75	1·2
0·95	1·4
1·10	1·7
1·20	2·0

(*e*) Select a relay, coil resistance R_L, which will operate reliably at a pull-in current I_{rel} such that

$$I_{rel} \leqslant \left[\frac{1 - r_2}{1 + r_1}\right]^2 \cdot \frac{I_{K(av)\,(max)}}{g} \qquad (5.89)$$

(*f*) Determine k.

$$k = \frac{V_m'\sqrt{2}}{V_M + g \cdot I_{rel} R_L} \qquad (5.77)$$

(*g*) From Fig. 5.7, determine the corresponding value of p.

(*h*) Determine the maximum nominal value of R_1,

$$R_{1(max)} = \frac{u(k-1)}{k \,.\, p} \, \frac{V_m'\sqrt{2}}{g \,.\, I_{rel}} \, \frac{(1 - r_2)}{(1 + w)} \qquad (5.79a)$$

If the anode circuit is according to Fig. 5.27 (*a*) or Fig. 5.30, adopt the nearest standard value below the calculated value of R_1.

(*j*) Calculate the peak current, $I_{K(pk)}$, corresponding to the worst case,

$$I_{K(pk)} \approx \frac{k-1}{k} \cdot \frac{V_m'\sqrt{2}}{R_1} \cdot \frac{(1 + r_1)}{(1 - w)} \qquad (5.90)$$

(*k*) Check that $I_{K(pk)} < I_{K(pk)\,(max)}$.

If this is not so, choose a relay of higher coil resistance. Alternatively, if a half-wave supply was proposed, consider using a full-wave supply.

(*l*) Check that $(u/p) \,.\, I_{K(pk)} \,.\, R_L < 90$

If this is not so, add D_2, R_5, and R_6 as in Fig. 5.30 (*b*), putting $R_5 = 1$ MΩ, $R_6 = 470$ kΩ.

(*m*) Note the percentage by which $I_{K(pk)}$ is below $I_{K(pk)\,(max)}$, or $(p/u)I_{K(av)\,(max)}$, or $90p/u \,.\, R_L$. Reduce the value of R_1 by half the smallest of these percentages to R_1'.

(*n*) Determine $\rho_{(max)} = V_m'/V_m$.

If $\rho_{(max)} < 1{\cdot}0$, the circuit of Fig. 5.27 (*b*) must be used. In this event, determine R_X' such that

$$R_X' \approx R_1'/\rho_{(max)} \qquad (5.91)$$

(*o*) Determine $\delta R_1/R_1$.

$$\delta R_1/R_1 = (R_1 - \rho_{(max)} \,.\, R_X')/R_1$$

(*p*) Determine the maximum permissible tolerance, t, on the value of R_X'.

$$t = \frac{1}{4} \cdot \frac{k-1}{k} \cdot \frac{1}{1 - \rho_{(max)}} \cdot \frac{\delta R_1}{R_1} \qquad (5.83a)$$

(*q*) Evaluate R_Y.

$$R_Y' = \frac{1-t}{1+t} \cdot R_X' \cdot \frac{\rho_{(max)}}{1 - \rho_{(max)}} \qquad (5.81c)$$

(*r*) Choose preferred values and tolerances for resistors such that R_X and R_Y always lie within the limits imposed on R_X' and R_Y' by the fractional tolerance, t.

(*s*) Choose C so that

$$C \approx 1/(2fu \,.\, R_L) \qquad (5.92)$$

(*t*) For shielded-anode tubes, determine from manufacturer's data the proportion, z, of V_m' to be applied to shield anode.

(*u*) Determine the approximate value of R_4.

$$R_4 \approx R_{A2}/z \qquad (5.86)$$

(*v*) Choose preferred values for R_3 and R_4 such that

$$\frac{R_3}{R_4} = \frac{z}{(1 - z)} \qquad (5.87)$$

In choosing R_3 and R_4, it should be remembered that the ratio, R_3/R_4, may be critical, whereas R_{A2} is not. In general, the nominal values of R_3 and R_4 should satisfy Equation (5.87) with an accuracy of better than 5% and 5% tolerance resistors should be used. If necessary, R_3 and R_4 may be reduced to obtain a required ratio and R_{A2} may be maintain by insertion of an additional resistor in series with A_2.

EXAMPLE 5.9 Self-resetting Relay Using Primed Triode

Design the anode circuit for a CV2434 self-resetting relay to operate on half-wave 50-c/s supply, nominally 220–240 V r.m.s. and subject to ±10% voltage variations.

(*a*) Set out tube data:

$V_{IG} = 290$ V $I_{K(av)\,(max)} = 25$ mA
$V_M = 105$ V $I_{K(pk)\,(max)} = 100$ mA
$V_{AT} = 170$ V

(*b*) Set out additional data:

$$220\text{–}240 \text{ V} \pm 10\% \approx 230 \text{ V} \pm 15\%$$

$$\therefore \; V_m = 230 \text{ V r.m.s. and } r_1 = r_2 = 0{\cdot}15$$

$$f = 50 \qquad u = \tfrac{1}{2}$$

Put $\quad w = 0{\cdot}10$

(*c*) $$V_m' \leqslant \frac{290}{1{\cdot}15\sqrt{2}} = \mathbf{178 \text{ V r.m.s.}}$$

(*d*) $$\frac{V_{AT}}{V_m'} = \frac{170}{178} = 0{\cdot}955$$

$\therefore\ g \approx 1{\cdot}5$, say.

(*e*) $$I_{rel} \leqslant \left(\frac{0{\cdot}85}{1{\cdot}15}\right)^2 \times \frac{25}{1{\cdot}5} = 9{\cdot}2\ \text{mA}$$

Hence choose a 2-kΩ relay for which $I_{rel} = 8$ mA.

(*f*) $$k = \frac{178\sqrt{2}}{105 + 1{\cdot}5 \times 8 \times 2} = \mathbf{1{\cdot}95}$$

(*g*) From Fig. 5.7, $p = 2{\cdot}0$

(*h*) $$R_{1(\max)} = \frac{\frac{1}{2} \times 0{\cdot}95}{1{\cdot}95 \times 2{\cdot}0} \times \frac{178\sqrt{2}}{1{\cdot}5 \times 8} \times \frac{0{\cdot}85}{1{\cdot}10} = \mathbf{1{\cdot}97\ k\Omega}$$

(*j*) In the worst case,
$$I_{K(pk)} = \frac{0{\cdot}95}{1{\cdot}95} \times \frac{178\sqrt{2}}{1{\cdot}97} \times \frac{0{\cdot}85}{1{\cdot}10} = \mathbf{80\ mA}$$

(*k*) $I_{K(pk)} < I_{K(pk)\ (\max)}$. Hence the design is practicable.

(*l*) $(u/p) \,.\, I_{K(pk)} \,.\, R_L = (\frac{1}{2}/2{\cdot}0) \times 80 \times 2 = 40$ V ($<$90 V)

Hence D_2, R_5, and R_6 are not required.

(*m*) $I_{K(pk)}$ is 20% below $I_{K(pk)\ (\max)}$ or $(p/u)I_{K(av)\ (\max)}$. Hence reduce R_1 by 10% to $R_1' = 1{\cdot}77$ kΩ.

(*n*) $$\rho_{(\max)} = 178/230 = \mathbf{0{\cdot}775}$$

Hence the circuit of Fig. 5.27 (*b*) must be used.

$$R_X' \approx 1{\cdot}77/0{\cdot}775 = 2{\cdot}29\ \text{k}\Omega,\ \text{say}\ \mathbf{2{\cdot}2\ k\Omega}$$

(*o*) $$\delta R_1/R_1 = (1{\cdot}97 - 0{\cdot}775 \times 2{\cdot}2)/1{\cdot}97 = \mathbf{0{\cdot}132}$$

(*p*) $$t = \frac{1}{4} \times \frac{0{\cdot}95}{1{\cdot}95} \times \frac{0{\cdot}132}{0{\cdot}225} = \mathbf{0{\cdot}071}$$

(*q*) $$R_Y' = \frac{0{\cdot}946}{1{\cdot}054} \times 2{\cdot}2 \times \frac{0{\cdot}775}{0{\cdot}225} = \mathbf{6{\cdot}8\ k\Omega}$$

(*r*) $\therefore\ R_X = 2{\cdot}2$ kΩ $\pm$7%,
$R_Y = 6{\cdot}8$ kΩ $\pm$7%.

(*s*) $$C \approx \frac{1}{2 \times 50 \times \frac{1}{2} \times 2{,}000} = \mathbf{10\ \mu F}$$

Solution:

Using the circuit of Fig. 5.27 (*b*),
$R_L = 2$ kΩ, operating at 8 mA,
$R_X = 2{\cdot}2$ kΩ $\pm$7%,
$R_Y = 6{\cdot}8$ kΩ $\pm$7%.
$C = 10\ \mu$F approx.

EXAMPLE 5.10 Self-resetting Relay Using Shielded-anode Tube

Design a self-resetting relay circuit for operation on 200–250 V r.m.s. 50-c/s supply with half-wave rectification. Use the Z806W shielded-anode tube.

(*a*) $V_{IG} = 390$ V $\qquad I_{K(av)\ (\max)} = 25$ mA
$V_M = 110$ V $\qquad I_{K(pk)\ (\max)} = 150$ mA
$V_{AT} = 220$ V $\qquad R_{A2} = 120$ kΩ

(*b*) $V_m = 225$ V r.m.s. $\qquad r_1 = r_2 = 0{\cdot}11$
$f = 50$ $\qquad u = \frac{1}{2}$

Put $w = 0{\cdot}10$

(*c*) $$V_m' \leqslant \frac{390}{1{\cdot}11\sqrt{2}} = 248\ \text{V r.m.s.}$$

Hence put $V_m' = V_m =$ **225 V r.m.s.**

(*d*) $$\frac{V_{AT}}{V_m'} = \frac{220}{225} = \mathbf{0{\cdot}98}$$

$\therefore\ g = 1{\cdot}5$, say.

(*e*) $$I_{rel} \leqslant \left(\frac{0{\cdot}89}{1{\cdot}11}\right)^2 \times \frac{25}{1{\cdot}5} = \mathbf{10{\cdot}7\ mA}$$

Select a relay, 4 kΩ coil resistance, operating at 5 mA.

(*f*) $$k = \frac{225\sqrt{2}}{110 + 1{\cdot}5 \times 5 \times 4} = \mathbf{2{\cdot}27}$$

(*g*) From Fig. 5.7, $p = 1{\cdot}95$.

(*h*) $$R_{1(\max)} = \frac{\frac{1}{2} \times 1{\cdot}27}{2{\cdot}27 \times 1{\cdot}95} \times \frac{225\sqrt{2}}{1{\cdot}5 \times 5} \times \frac{0{\cdot}89}{1{\cdot}10} = \mathbf{4{\cdot}92\ k\Omega}$$

(*j*) In the worst case,
$$I_{K(pk)} = \frac{1{\cdot}27}{2{\cdot}27} \times \frac{225\sqrt{2}}{4{\cdot}92} \times \frac{1{\cdot}11}{0{\cdot}90} = \mathbf{44{\cdot}5\ mA}$$

(*k*) $I_{K(pk)} < I_{K(pk)\ (\max)}$. Hence the design is practicable.

(*l*) $(u/p) \,.\, I_{K(pk)} \,.\, R_L = (\frac{1}{2}/1{\cdot}95) \times 44{\cdot}5 \times 4 = 45{\cdot}5$ V ($<$90 V)

(*m*) $I_{K(pk)}$ is 70% below $I_{K(pk)\ (\max)}$,
54% below $(p/u) \,.\, I_{K(av)\ (\max)}$,
and 49% below $90p/u \,.\, R_L$.

Hence reduce R_1 by approximately $\frac{1}{2} \times 49$%, i.e. by 24·5%.
$\therefore\ R_1' \approx (1 - 0{\cdot}245) \times 4{\cdot}92 = 3{\cdot}72$ kΩ, say **3·9 kΩ**

(*n*) From (*c*), $\rho_{(\max)} = 225/225 = 1{\cdot}0$
Hence the circuit of Fig. 5.30 (*a*) is acceptable.

(*o*), (*p*), (*q*), and (*r*) do not apply.

(*s*) $$C \approx \frac{1}{2 \times 50 \times \frac{1}{2} \times 4{,}000} = 5\ \mu\text{F}$$

(*t*) $$z \approx 0{\cdot}7$$

(*u*) $$R_4 = 0{\cdot}7 \times 120 = 84\ \text{k}\Omega\text{, say } \mathbf{82\ k\Omega}.$$

(*v*) $$\frac{R_3}{R_4} = \frac{0{\cdot}7}{0{\cdot}3}$$

$$\therefore\ R_3 = 200\ \text{k}\Omega\ \pm 5\%$$

$$R_4 = 82\ \text{k}\Omega\ \pm 5\%.$$

Solution:

In the circuit of Fig. 5.30 (*a*),

$R_L = 4$ kΩ, operating at 5 mA,
$R_1 = 3{\cdot}9$ kΩ $\pm 10\%$,
$R_3 = 200$ kΩ $\pm 5\%$,
$R_4 = 82$ kΩ $\pm 5\%$,
$C = 5$ μF approx.

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CHAPTER SIX

Arc Discharge Tubes

If a cold cathode tube is connected directly in parallel with a charged capacitor, breakdown of the anode–cathode gap allows the discharge to carry a current which is limited only by the 'internal resistance' of the discharge and by resistance and inductance of the leads from capacitor to tube or in the capacitor itself. It follows that, unless the capacitance is kept so small that it discharges before the tube has become heavily ionized, the discharge will pass rapidly through the normal and abnormal glow regions to become an arc (Fig. 2.2). The discharge then centres on a small spot on the cathode and the high current density (100–1,000 A/cm^2) leads to rapid destruction of a cathode not suited to this mode of operation.

Three classes of cold cathode arc tubes are discussed here:

(1) Stroboscopic flash and switching tubes provided with internal electrodes permitting triggering from low-voltage, low-energy pulses.

(2) Photographic and stroboscopic flash tubes and switching tubes in the form of 'diodes' with an externally wrapped trigger lead requiring a high-voltage triggering pulse.

(3) Thyratrons ('Arcotrons') designed to carry continuous currents of several Amperes.

Arc Discharge Tetrodes

In 1928 Steinert [1] described the first use of a neon tube to provide a high-resolution stroboscope with no moving parts. A saturated-core transformer connected to the supply mains generated a peaked waveform which was applied to a neon diode and 'tuning capacitor' connected in shunt across the secondary winding. On each half-cycle of the supply the lamp produced a flash of duration estimated at 20 μsec.

Four years later, Quarles [2] described a neon-flash stroboscope with many features characteristic of modern devices, but using the KU-610 hot cathode tube. This grid-controlled tube provided a 0·3-μsec flash in response to triggering pulses from a multivibrator or electrical contacts.

The work of Edgerton, Germeshausen, and Grier led to the cold

cathode neon tetrode exemplified in the Sylvania 'Strobotron' and Ferranti 'Neostron'. Hilliard [3] describes the Sylvania 1D21/631P1 illustrated in Fig. 6.1. Between anode and cathode lie a cylindrical

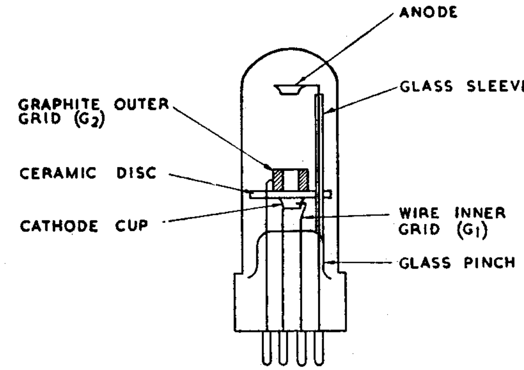

FIG. 6.1. Electrode structure of arc tetrode (Sylvania 1D21/631P1).

graphite outer grid, G_2, and, nearer the cathode, a wire probe inner grid, G_1. The control characteristics of the tube are complicated by the number of variables involved: given suitable applied voltage, a discharge

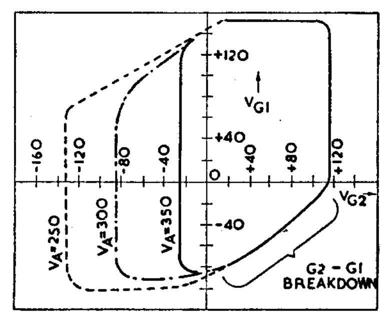

Fig. 6.2. Ignition characteristic of arc tetrode. Triggering is usually effected by breakdown between G_2 and G_1.

may be initiated in either direction between any pair of electrodes. A study of the triggering characteristics in Fig. 6.2, however, shows that the anode voltage has little effect on the critical grid potentials for the normal mode of operation. This comprises holding G_2 positive and applying a negative-going pulse to G_1. A pulse of -30 V will suffice

with a positive bias of about 100 V on G_2. The triggering characteristics tend to change during the life of the tube, however, and it is therefore preferable to bias less critically and apply a much larger trigger pulse. If this is generated by a trigger tube operating in the self-quenching mode this may easily be of the order of 100 V.

With a negative-going pulse on G_1, breakdown between anode and cathode follows after a formative delay which, though dependent on anode voltage and, to some extent, on pulse amplitude, is independent of pulse duration. Feinberg [4] has studied the performance of neon-filled and argon-filled tubes with pulses of both polarities. With positive pulses on G_1 he finds anode breakdown tends to occur at a fixed time after the *termination* of the trigger pulse. This is not surprising, since with positive pulses the field between G_1 and G_2 is in a direction to oppose the diffusion of electrons into the anode–G_2 gap. With increasing anode voltage this relation expires at progressively shorter pulse durations. Feinberg was able to produce a controlled delay of 10–400 μsec.

Once an anode–cathode breakdown is established, it is important that the cathode current be allowed to rise rapidly to a sufficiently high value ($\sim$5 A) to produce an intense cathode spot and so allow the discharge to become an arc. To this end, a cathode is used containing caesium. The capacitor discharges rapidly through the arc until a charge of only 20 V remains. At this voltage the discharge normally extinguishes. As the arc impedance is low, excessive series inductance – due to long leads, for example – can lead to voltage backswing and inverse conduction. The anode then assumes the role of cathode, and severe sputtering may result. This will produce blackening of the envelope and some clean-up of the gas. Series inductance should therefore be minimized (e.g. by shortening and twisting leads) or, if it

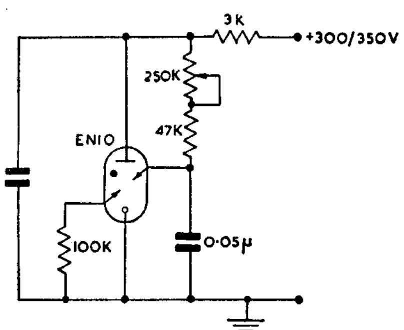

Fig. 6.3. Free-running stroboscope.

is unavoidable, sufficient circuit resistance should be introduced to provide critical damping.

A simple free-running stroboscope may be constructed to the circuit of Fig. 6.3. In operation such a stroboscope may show some 'jitter' due to its sensitivity to changes in supply voltage and tube characteristics. It is accordingly preferable to synchronize the stroboscope tube by applying negative-going pulses to G_1 from a separate multivibrator or blocking oscillator.

In addition to the neon-filled tubes, similar tubes are available with other gas-fillings (e.g. argon) better suited to photographic applications.

Switching Diodes and Flash Tubes

Edgerton, Germeshausen, and Grier were responsible also for the first photographic flash tubes of the 'electronic' variety, as distinct from the

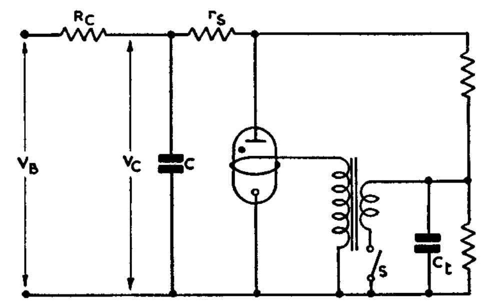

Fig. 6.4. Simple flash tube circuit.

expendable incandescent flash bulbs. The photographic flash tubes were based on the same research work as the tetrodes, but were designed specifically for high actinic light output. Instead of neon, argon or xenon–krypton fillings are used to provide an almost white flash. The discharge path is also lengthened to increase the parameter which Murphy and Edgerton [5] call 'tube resistance'. Finally, the tube is designed to accept a very high instantaneous power dissipation ($\sim$1MW).

Fig. 6.4 shows a simple circuit for the operation of a flash tube. Closure of switch S discharges C_t through the primary of the triggering coil. Capacitive coupling of the high-voltage impulse applied to the external trigger ($\sim$10 kV) is sufficient to distort the potential gradient within the tube and lead to breakdown. The storage capacitor, C, then discharges through the tube, the current being limited by the 'tube resistance', r_t, and the series resistance, r_s, of the leads.

The energy, E, stored in the capacitor is given (in Joules, or watt-seconds) by:

$$E = \tfrac{1}{2}CV_C^2 \qquad (6.1)$$

If the arc extinguishes at a voltage, v, the energy, e, remaining in the capacitor is given by:

$$e = \tfrac{1}{2}Cv^2 \qquad (6.1a)$$

The energy drawn from the capacitor is thus

$$(E - e) = \tfrac{1}{2}C(V_C^2 - v^2) \qquad (6.2)$$

Some of this is dissipated in series resistance, r_s, the rest in the tube. It is the designer's object to maximize one of these dissipations and minimize the other. In a switching tube he attempts to minimize the tube losses. If, to this end, the 'tube resistance' can be made negligibly small the energy dissipated by the tube is given by:

$$E_t = (Q - q)v = C(V_c - v)v \qquad (6.3)$$

where Q = initial charge on C,

and q = final charge on C.

The efficiency of energy transfer from capacitor to *tube* is then given by η_0, where

$$\eta_0 = E_t/(E - e)$$

$$\therefore \eta_0 = 2v/(V_C + v) \qquad (6.4)$$

In a typical tube, $v \approx 20$ V. Switching tubes operate with anode voltages as high as 10 kV. With negligible internal impedance, such tubes should therefore give practical efficiencies of $>99\%$, i.e. this percentage of the stored energy is delivered into the *load.*

In a photographic flash tube it is the object to transfer as much energy as possible into the tube. These tubes are therefore made with long discharge paths, often folded into a U or a helix. This produces a significant voltage drop in the positive column, and Murphy and Edgerton [5] have shown that, once a discharge has been initiated, such a tube behaves substantially as a resistance of 1–3 Ω.

Extending the foregoing treatment to cover a tube comprising a positive column of effective resistance, r_t, in series with a cathode fall, v,

$$\eta = E_t/(E - e)$$

$$= 1 - \frac{E_s}{(E - e)} \qquad (6.4a)$$

where E_s is the energy dissipated in the series resistance, r_s, of the external circuit.

Now

$$E_s = \tfrac{1}{2}(Q - q)V_s$$

$$= \tfrac{1}{2}[C(V_C - v)]\left[\frac{r_s}{r_t - r_s} \cdot (V_C - v)\right] \qquad (6.5)$$

$$\therefore \eta = 1 - \left(\frac{r_s}{r_t + r_s}\right)\left(\frac{V_C - v}{V_C + v}\right) \qquad (6.6)$$

From Equation (6.6) it will be seen that, to ensure a high efficiency, η, of energy transfer from capacitor to tube, it is necessary that $r_t \gg r_s$. As it is not easy to keep r_s much below 0·1 Ω, the length of a photographic flash tube is usually made sufficient to raise r_t to 2 or 3 Ω. The value of r_t varies almost in proportion to tube length.

Light Output

The nature and efficiency of light emission by a flash tube has been studied by Tuttle, Brown, and Whitmore [6] and by Olsen and Huxford [7].

Tuttle and others remark that, with tubes having an effective resistance of 1 Ω, the flash duration (in seconds) is equal to the storage capacitance (in Farads). Very short flash durations, of the order of 1 μsec, are achieved only if great care is exercised to minimize resistance and inductance in the leads connecting tube and capacitor. Flash durations of 30–100 μsec are more typical for tubes operating at about 2 kV and about 1 msec for low-voltage tubes. The majority of photographic flash tubes have a tube resistance of about 3 Ω. Hence, for such a tube, the flash duration (in seconds) is about three times the value of C (in Farads).

In the tubes studied by Olsen and Huxford [7] the tube current and power dissipation rose to a peak at about 2 μsec from breakdown. Thereafter the capacitor discharged exponentially according to the relation

$$i = I_{(\max)} \,.\, e^{-t/Cr_t} \qquad (6.7)$$

The curve of light emission against time is less sharply peaked than the i/t curve and reaches a peak 4–8 μsec after the peak current occurs. The spectral distribution of the light emission changes during the flash, containing a preponderance of ultra-violet at first and infra-red at the end of the flash. For most practical purposes, however, the flash may be considered white and matching daylight. The high loading of the arc broadens the lines characteristic of the gas used and produces such a large continuum that they are no longer significant.

When $V_C \gg v$ it may be deduced from Equations (6.2) and (6.6) that the light output, $\mathscr{L}$, will be governed by the relation:

$$\mathscr{L} \propto C \,.\, V_C^2 \qquad (6.8)$$

In practice, the light output increases more rapidly than either C or V_C^2. This effect is more pronounced at shorter wavelengths and is apparently due to the higher level of ionization produced by higher-energy discharges.

In the more general form of Relation (6.8), viz. $\mathscr{L} \propto (C \,.\, V_C^2)^n$, Olsen and Huxford report that for the infra-red band, $n = 1$ for xenon, 1·2 for argon, and 1·6 for neon. In the visible band $n = 1{\cdot}8$ for argon and 2·1 for neon. In the ultra-violet $n = 2{\cdot}3$ and 2·8 for the same two gases.

Provided C and V_C remain constant, a flash tube provides excellent consistency of light output. V_C may conveniently be stabilized by a trigger tube shunt stabilizer of the type shown in Fig. 5.4. For the most precise work, the first flash of each series should be ignored as the light output is slightly lower than that produced by subsequent flashes.

With a single flash an efficiency of 50 lumens/watt may be achieved. Stroboscopic flash tubes studied by Carlson and Edgerton [8] produced only 20–30 lumens/watt due to the reduced loading which could be applied at each discharge.

Limits of Loading

With single flashes the tube loading is usually restricted by heating of the walls. Excessive loading produces 'crazing' of the glass which permits out-gassing. This in turn leads to a rise in breakdown voltage so that the tube becomes difficult to trigger. To some extent this may be offset by extending the length of trigger lead wrapped around the tube envelope. (Conversely, if a tube fires spontaneously the wrap-round of trigger lead may often be reduced with advantage.)

When a tube is flashed repetitively, as in stroboscopic applications, the mean dissipated power, $P_{(av)\,(max)}$, is important. Glass will conduct if heated excessively, and the trigger impulse may then puncture the tube wall. Glass tubes may also fail due to softening. Quartz softens at a much higher temperature, and is therefore used for high-power stroboscope tubes. Repetitive flashing also increases the electrode temperatures and stroboscope tubes must accordingly use relatively large electrodes capable of dissipating the heat developed. Carlson and Edgerton [8] describe experiments with tubes operating at a mean dissipation of 5 kW with forced air cooling.

If the maximum average power dissipation, $P_{(av)\,(max)}$, for a given

cooling system is known the maximum energy, E, per flash (assuming 100% efficiency of power transfer to tube) is given by:

$$E = P_{(av)\,(max)}/f = \tfrac{1}{2}CV_C^2 \qquad (6.1b)$$

As the frequency, f, of flashing is raised, the energy per flash must be proportionately reduced. From the foregoing, it will be appreciated that this reduces the efficiency of light generation, particularly at shorter wavelengths.

Charging Circuit Design

Referring to Fig. 6.4, the theoretical maximum efficiency of 50% is approached if $CR \leqslant 1/4f$. This means that the mean power dissipated in R_C is the same as that in the discharge circuit. Thus, since C has charged to >98% of V_B,

$$\text{Watts}_{R_C} \approx \tfrac{1}{2}fCV_C^2 \approx \tfrac{1}{2}fCV_B^2 \qquad (6.9)$$

Hilliard [3] provides design curves useful when $CR > 1/4f$.

For stroboscopic applications in which power economy is important, the charging resistor, R_c, may be replaced by a choke and diode, as in Fig. 6.5. After the first discharge the choke swings positive to charge

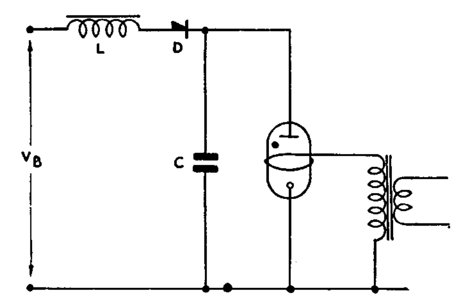

Fig. 6.5. Stroboscopic flash tube with high-efficiency power supply.

C to $2V_B$. The diode D then prevents loss of charge. This arrangement approaches 100% electrical efficiency and, moreover, provides voltage doubling so that the value of C may be reduced to one-quarter for a given energy per flash. Carlson and Edgerton [8] discuss the design and operation of this type of charging circuit. They summarize their findings in three recommendations:

(1) The peak energy-storage capacity of the inductor, L, must be at least one-fourth the final energy storage of the capacitor.

i.e. $$L \geqslant \tfrac{1}{4}[\tfrac{1}{2}CV_C^2] \qquad (6.10)$$

(2) For a given capacitance, the inductance should not be so large that $\pi\sqrt{LC}$ exceeds the period between flashes.

i.e.
$$L < \frac{1}{(\pi f)^2} \cdot \frac{1}{C} \tag{6.11}$$

(3) The inductance should not be so small that recharging is effected much more rapidly than required by the frequency of flashing. A low inductance increases the danger of 'hold-over' due to V_C rising to restriking potential before deionisation is completed.

Cold Cathode Arc Thyratrons ('Arcotrons')

Whereas the tubes so far described in this chapter are all designed to carry large or very large currents for short periods (~1,000 A for ~10 μsec) many applications call for a thyratron capable of carrying a few Amperes continuously. In 1961 the Swiss firm Cerberus, of Maennedorf, announced a range of cold cathode thyratrons representing an attractive alternative to both hot cathode thyratrons and silicon-controlled rectifiers. The standby power consumption (<0·5 W) of these tubes is much less than that of comparable hot cathode thyratrons. In 1961 the overload capacity and price of the cold cathode devices compared very favourably with solid state equivalents. On the other hand, the arc drop (~20 V) restricted the value of the 'Arcotron' to applications in which the consequent power loss was not serious.

Fig. 6.6 shows diagrammatically the 'Arcotron' as described by

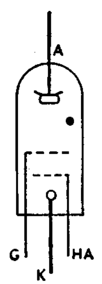

Fig. 6.6. Diagrammatic representation of 'Arcotron' tube.

Seifert [9]. A priming current of 10 or 20 mA flows from an auxiliary anode, HA, to the cathode K. The arc drop between these two electrodes is 20 to 25 V. Some of the electrons accelerated towards the auxiliary anode pass through perforations in HA to enter the space between HA and a control grid, G, which is also perforated. If G is held negative its retarding field repels electrons and they return to the auxiliary anode. On the other hand, if the potential of G is zero or positive electrons are accelerated by the positive main anode, A, and, entering the gap between

G and A, they produce breakdown. Once an arc is established between G and A in this way it may be interrupted only be reducing V_{A-K} below the arc maintaining voltage of ~20 V.

Because the field of the main anode extends through the grid perforations, the critical grid potential depends to some extent on the applied anode voltage. Nevertheless, Fig. 6.7 shows that, for a Type

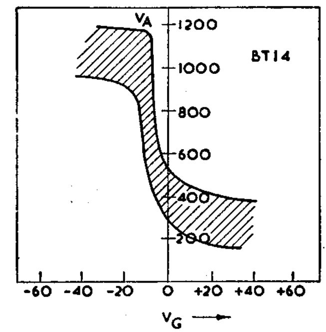

Fig. 6.7. Control characteristic of BT14 'Arcotron'.

BT14 tube, operating with anode voltages in the range 600–800 V, a grid bias of −15 V will always inhibit firing and the tube will always fire if the grid becomes more positive than −2 V. Control by transistors is therefore quite practical.

The 'Arcotron' should not be confused with another Cerberus tube which has been described by Pun and Mitter [10]: the glow thyratron (p. 65). In the latter tube the discharge operates in the normal glow region and the maximum continuous current is accordingly much smaller (~40 mA). 'Arcotrons', on the other hand, can carry continuous currents of 3 or 6 A and peak currents of 100 or 200 A.

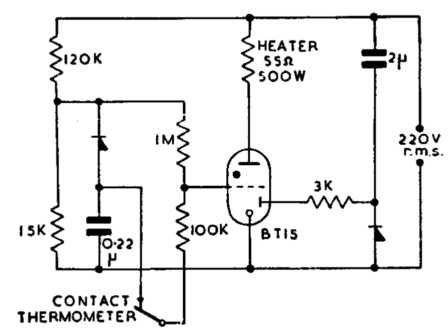

Fig. 6.8 Circuit for thermostatic control using BT15 'Arcotron'.

Fig. 6.8 represents the application of a Type BT15 tube to the thermostatic control of a water bath. Only a few microamperes flowing through the electrodes of the contact thermometer initiate switching of 3 A r.m.s. in the load. The tube deionizes on each negative half-cycle of the a.c. supply. On each positive half-cycle, the auxiliary arc restrikes and the main arc is established if the thermometer contacts do not apply a negative bias to the grid.

Other tubes in the range are capable of switching continuous loads of up to 2 kW.

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CHAPTER SEVEN

Stepping Tubes

The proprietary name 'Dekatron' has become almost synonymous with the cold cathode stepping tube. Since the introduction of the first Ericsson counting tubes in 1949, however, comparable tubes have been made available by other manufacturers, and today the stepping tube assumes a variety of forms. That originally described by Bacon and Pollard [1] has survived almost unchanged and, with the slightly later form shown in Fig. 7.1, is still widely used.

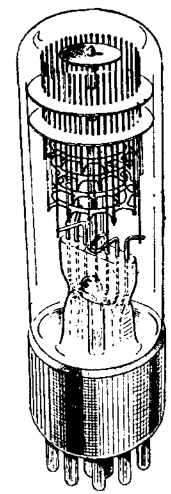


Fig. 7.1. The GC10D 'Dekatron' stepping tube.

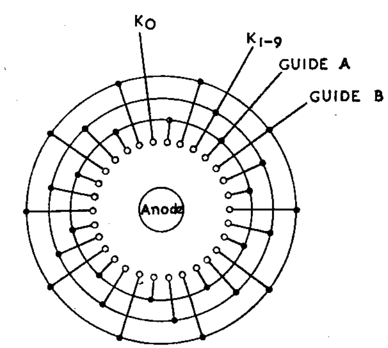


Fig. 7.2. Electrode arrangement of stepping tube with 30 cathodes.

In its most common form the stepping tube comprises thirty rod-like nickel cathodes disposed in a ring around a single disc-shaped anode. The whole is enclosed in a gas-filled glass envelope and is provided with a multi-pin base. Every third cathode, K_0, K_1, . . . K_9, in Figs. 7.2 and 7.5 (*a*), is an 'index cathode' while the intermediate electrodes are known as 'guides'. (Often the index cathodes are referred to merely as

'cathodes'.) The guide adjacent to an index cathode and displaced from it in a clockwise direction is called 'Guide A'. Between Guide A and the next index cathode lies 'Guide B'. All 'A' guides are connected

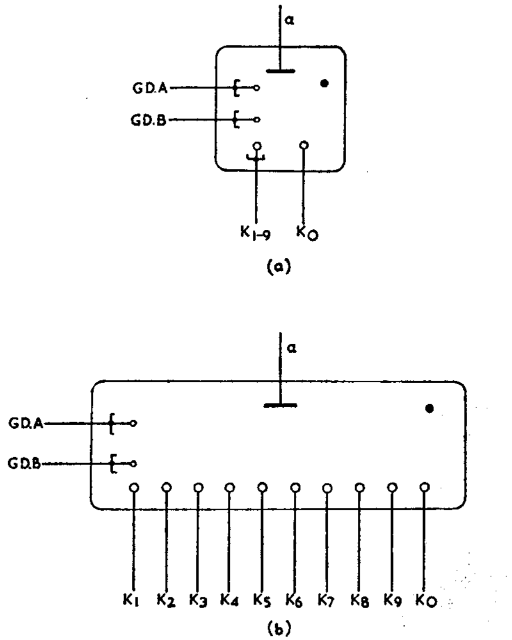


Fig. 7.3. Symbolic representations of stepping tubes: (*a*) counter with common connexion to nine index cathodes and separate connexion to K_0, (*b*) selector tube with separate connexion to each index cathode.

together, and all 'B' guides are also connected. Nine of the index cathodes are also connected. One remains separate so that an output pulse may be generated on every tenth count. Thus only five external connexions may be provided, as follows:

(1) Anode
(2) Guide A
(3) Guide B
(4) Common cathodes
(5) Output cathode

The symbolic representation of this tube is shown in Fig. 7.3 (*a*). Fig. 7.3 (*b*) represents a selector tube in which each index cathode is brought to a separate pin so that thirteen connexions are required.

Principle of Operation

A bias of about +40 V is applied to both sets of guides. An h.t. voltage of about 475–500 V is applied to the anode through an anode load resis-

tor of about 330–820 kΩ. Natural radiation provides sufficient primary ionization to cause breakdown to occur, although in the dark this may take some minutes. When a discharge takes place between anode and one index cathode it is visible through the end of the tube as a glow on the cathode concerned. A numbered escutcheon is commonly placed around the tube so that the glowing cathode may readily be identified.

In this static condition the anode–cathode maintaining voltage, V_M, is typically 190 V. The position of the discharge is perfectly stable, since the guides on either side of the conducting cathode are biased positively and the discharge tends always to seek a more negative electrode in its vicinity.

This fact is used to transfer the discharge from one cathode to the next when a count is to be registered. A pulse is applied to Guide A to drive it below cathode potential. As a result, the glow discharge transfers to the adjacent Guide A electrode. Before the Guide A potential has again risen a pulse is then applied to Guide B so that this also is depressed below cathode potential. As the Guide A potential rises, therefore, the glow transfers to the adjacent Guide B electrode. Finally, as the Guide B potential rises once more to the bias of +40 V the glow transfers to the adjacent index cathode. Here it rests until the appropriate pulse sequence is again applied to the two guides.

At each stage the discharge is moved by offering it an adjacent new cathode at a lower potential. The breakdown voltage of this adjacent anode–cathode gap is reduced to about 200 V by the neighbouring glow discharge. The discharge readily moves into this gap if the new cathode is made as little as 10 V negative with respect to the cathode on which the discharge rests. Once the new guide electrode has received the discharge the anode assumes a potential 190 V above it. This is too low to maintain a discharge to the former cathode, and so only the new guide electrode is left conducting.

When the discharge rests on, say, K_4, the breakdown voltage from anode to GDA_5 is about 200 V, but to GDA_4 and GDA_6 about 250 V. This is because these electrodes are farther away from the glow and so less heavily primed. Thus, although all Guide A electrodes are pulsed simultaneously to the same potential, there is normally no danger that the discharge will transfer to other than an adjacent electrode.

By connecting a resistor in series with the K_0 cathode, an output pulse of about 20 V may be obtained which can operate a second counter through a suitable pulse amplifier.

An important advantage of the type of tube discussed so far lies in its ability to step in either direction. Reverse counting requires only

reversal of the sequence of pulses applied to Guides A and B. Some complication of inter-stage coupling is involved, however.

Types of Guide Pulse

Guide pulses may assume three forms:

(*a*) Staggered double rectangular pulses showing some degree of overlap, as in Fig. 7.4 (*a*). Of the three methods, this usually requires

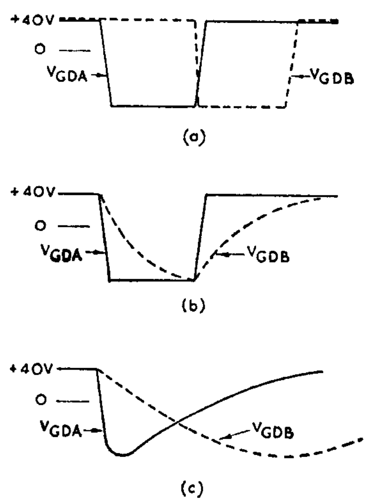


Fig. 7.4. Types of guide pulse: (*a*) double rectangular, (*b*) integrated pulse, (*c*) pulse differentiated at Guide A, integrated at Guide B.

the most complex circuitry for pulse generation. On the other hand, since both the leading and trailing edges of the pulses are well defined, the glow rests on each cathode for a controllable and well-defined period. This may be a valuable feature when an output is taken from one or more cathodes to operate other circuits, particularly when logical circuits are involved. Chaplin [35] reports that, when tubes are to be operated at maximum speed, this method of drive provides the greatest reliability of stepping.

The probe of a revolution counter can often be arranged to generate two pulses in an appropriate displacement. Reversibility is then obtained directly as the phasing of the pulses reverses with the direction of rotation.

(*b*) Integrated pulse drive, as indicated by Fig. 7.4 (*b*). A rectangular pulse is applied to Guide A and a simple *RC* integrator applies a delayed pulse to Guide B.

This type of drive is simpler to provide and does not entail the danger of a space between pulses. It can be made to operate satisfactorily at high speeds, but as the trailing edge of the Guide B pulse is not well defined, the leading edge of an output pulse from an index cathode will show considerable 'jitter'. Only the trailing edge of the cathode output pulse is well defined.

(*c*) A combination of differentiated pulse on Guide A and an integrated form of the same pulse on Guide B. This is shown in Fig. 7.4 (*c*).

If the time constants are chosen to hold Guide A below Guide B long enough to ensure that it accepts the discharge, then the long exponential recoveries of the guide potentials severely restrict the maximum speed of operation. This may not be of consequence in the later stages of a decade scaler, and the circuit is frequently used in this application.

Other Types of Stepping Tube

In addition to those tubes described above, similar tubes are available having 12 index cathodes and 24 guides. These effect glow transfer by the methods described already. So also do the tubes provided with auxiliary anodes (p. 203) and the low-voltage stepping tube (p. 195) designed for transistor drive.

Various alternative forms of stepping tube have been described [2, 3, 4] and are indicated diagrammatically in Figs. 7.5 (*b*), (*c*), and (*d*). These may be compared with the type already discussed and indicated by Fig. 7.5 (*a*). All are based on the principle of extinguishing the discharge to one cathode by initiating a discharge to an adjacent, temporarily more negative, cathode.

An important example is the single-pulse Dekatron (Fig. 7.5 (*b*)), of which the GC10D was the first type commercially available. This tube remains substantially as first described by Acton [3] in 1952, although developments have been made in the associated circuitry.

The GC10D differs from the double-pulse 'Dekatron' in having 40 cathodes in place of the usual 30. Also a different gas filling is used. Between each pair of index cathodes there are three guide electrodes: Guide X, Guide Y, and Guide Z. All Guide X connexions are common, as also are all Guide Y connexions. All Guide Z electrodes are also common with the exception of the one next to the output cathode K_0. Figs. 7.5 (*b*) and 7.6 show that the Guide X connexion is returned to Guide Y through 220 kΩ shunted by 100 pF. The common Guide X is similarly returned to the common index cathode connexion, K_{1-9}, and the output Guide Z_0 in like manner to K_0. In the static condition

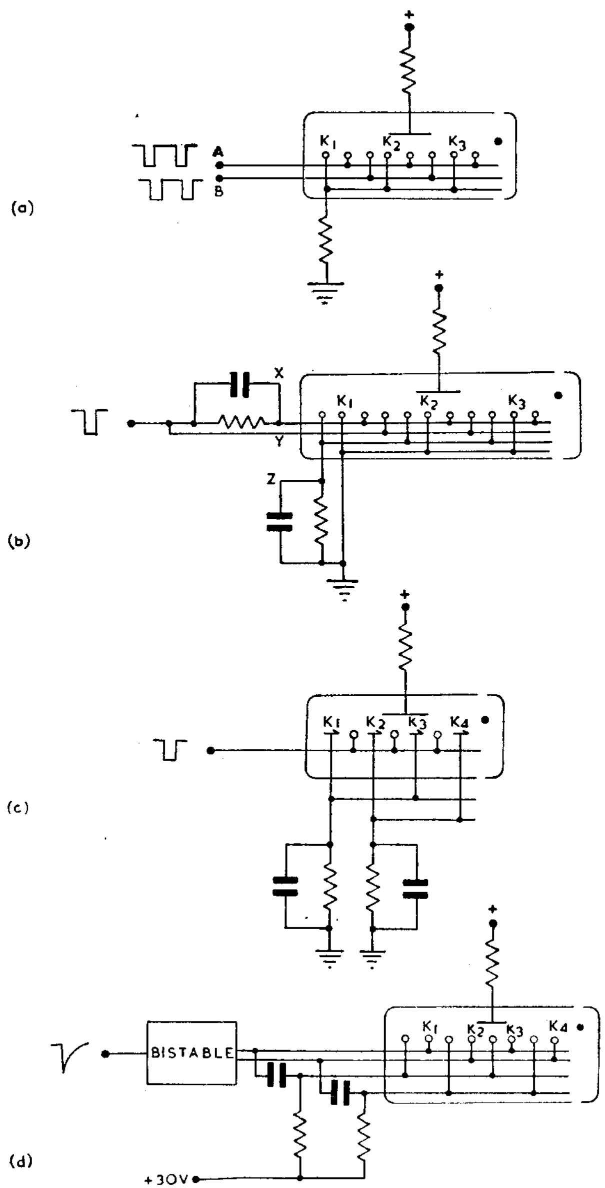


Fig. 7.5. Alternative types of stepping tube: (*a*) double-pulse tube, (*b*) single-pulse 'Dekatron', (*c*) single-pulse tube in which direction of stepping is determined by electrode geometry, (*d*) high-speed tube using grouped 'odd' and 'even' index cathode and correspondingly grouped guides.

Guides X and Y are biased to about +70 V. When the anode is returned to more than +420 V a discharge occurs to one index cathode or to one of the Guide Z electrodes. If the latter occurs, the 100-pF capacitor in the Guide Z return lead allows the discharge to become established before the *IR* drop in the 220 kΩ resistor causes the Guide Z potential to rise significantly. As the capacitor charges, however, the Guide Z potential rises and the discharge transfers automatically to the adjacent index cathode. Thus the static condition corresponds always to that in which the glow rests on an index cathode, any tendency to depart being opposed, on one side, by the +70 V on Guide X and, on the other, by the rise in Guide Z potential when Guide Z carries current.

To register a count, a negative-going pulse is applied to Guide Y carrying it to about −75 V. As Guide X is returned (through 220 kΩ shunted by 100 pF) to Guide Y, Guide X is carried negative with Guide Y. As a result, the glow discharge which was resting on an index cathode transfers to the adjacent Guide X electrode. Almost immediately the 100 pF in the Guide X lead is charged so that Guide X is positive with respect to Guide Y although still negative with respect to the index cathode. Accordingly, the discharge transfers automatically to Guide Y. No matter how long the pulse on Guide Y, the *CR* combination prevents the glow from returning to Guide X. When Guide Y returns to the normal positive bias it becomes more positive than Guide Z. Accordingly, the discharge transfers to Guide Z and thence to the succeeding index cathode, as described previously.

It will thus be seen that the GC10D requires only a single driving pulse and, provided this is of sufficient duration (>25 μsec), its shape is not critical. Moreover, the choice of gas-filling provides a relatively short de-ionization time. The GC10D can thus operate at stepping frequencies up to 20 kc/s. On the other hand, it is not readily used for reversible counting.

Fig. 7.5 (*c*) represents a type of single-pulse tube in which the direction of stepping is determined by electrode geometry. Hough and Ridler [2] have described tubes made by Standard Telephones and Cables Ltd. in which the glow is received by the edge of an index cathode, but automatically moves to the plate area, thereby priming the next guide cathode corresponding to clockwise rotation. A similar principle is used in the high-speed stepping tubes described by Apel [4] and manufactured by Elesta A/G. In these a central anode is surrounded by twenty cathodes of hooked form. The cathodes are tilted from a radial disposition so that the hooked inner end of each cathode is advanced in the direction in which the glow is to step. In the static

condition the glow rests on the rounded inner end of a hooked index cathode. The tail of the next guide cathode acts as a probe in the discharge, and accordingly is heavily primed by it. When this guide is pulsed negatively, therefore, it immediately accepts the discharge and the glow moves to the rounded end of the hook, thereby priming the next index cathode. As the guide potential rises at the end of the pulse, the discharge is therefore accepted by the primed index cathode. The EZ10 can operate at up to 100 kc/s and the EZ10B at 1 Mc/s.

In both the S.T. & C. and the Elesta tubes it is important that adjacent index cathodes shall have separate cathode resistors and by-pass capacitors. At the end of a short guide pulse the residual bias on the previously conducting cathode then inhibits restriking of the still ionized gap. In many cases it is sufficient to use only two cathode *RC* combinations, one connected to all odd cathodes and the other to all even cathodes, as in Fig. 7.5 (*c*).

Exceptional reliability of stepping is claimed for the Elesta ECT100, illustrated diagrammatically in Fig. 7.5 (*d*). Twenty cathodes are used, comprising the tips of four star-shaped stampings insulated from each other, but mounted co-axially within a cylindrical anode ring. The arms are bent slightly so that their tips all lie in the same plane. Separate leads are brought out from each stamping, one representing all the odd-numbered index cathodes, another all the even. The other two stampings are guide electrodes, G1 and G2, which are *RC* coupled to the 'odd' and 'even' cathodes respectively. The 'odd' and 'even' cathodes are coupled directly to a bistable circuit so that one is at ground potential, the other at +60 V. In a quiescent condition G1 and G2 were biased to +30 V.

Suppose the discharge rests on K_1, which is at ground potential. On receipt of an input pulse the bistable reverses state. K_1 rises to +60 V and K_2 falls to ground potential. Through the *RC* coupling, G1 is driven momentarily to >60 V and G2 to below ground. The discharge therefore transfers from K_1 to the adjacent G2 electrode. Then, as G2 rises once more to +30 V, the discharge transfers to K_2.

The tube is provided with output probes close to the tips of the ten index cathodes. By ionic coupling an output of up to 200 V may be obtained from the probe opposite the conducting index cathode.

It is claimed that, besides operating at up to 1 Mc/s under severe environmental conditions, the ECT100 shows virtual immunity from 'sticking' (p. 183) following prolonged operation with the discharge resting on a single index cathode. The tube will also operate with

unusually large supply voltage variations, and h.t. stabilization is not generally needed. The relatively small drive pulse amplitude is readily provided by a pair of transistors. As the electrode arrangement is symmetrical, the tube may be stepped in reverse if the *RC* couplings to G1 and G2 are interchanged.

Guide Pulse Dimensions

The glow begins to transfer from one electrode to the next when the adjacent electrode reaches a potential 10 V below that of the electrode on which the glow is resting. Complete transfer will not occur unless the potential difference reaches 20 V. With a neon-filled tube the transfer time is then about 100 μsec. This time is reduced if the potential difference is further increased. With a new tube a 60-V difference may reduce the transfer time to 25 μsec, but as the tube ages this will become asymptotic to a value of about 75 μsec.

It is useful to apply the name *guide transfer voltage* (V_{GT}) to that part of the guide pulse amplitude which is effective in producing transfer from one electrode to another. Thus the guide transfer voltage is the voltage by which a guide is made more negative than either of its neighbours. The effective pulse duration is that time for which the guide is held more negative than its neighbours. With double rectangular pulses (Fig. 7.4 (*a*)) these values are rarely in doubt. With other shapes of drive pulse some care must be exercised. In Fig. 7.4 (*c*), for example, the effective duration of the Guide A pulse is the time t_A during which Guide A is more negative than Guide B. The effective transfer voltage on Guide A is more difficult to determine. The instantaneous value of guide transfer voltage falls as the potential of Guide B falls. Over the time t_A, therefore, the effective pulse amplitude is less than its peak amplitude. Integrated pulse drive (Fig. 7.4 (*b*)) commonly requires a large value of t_A, even though the peak value of V_{GT} may be made larger than that used with double integrated pulses.

A maximum value of guide pulse is set by the danger of breakdown between any cathode and a negatively pulsed guide. If such breakdown occurs, current will flow from one electrode to the other. The electrode which then acts as an anode will have its working surface destroyed, and the tube will thereafter refuse to step correctly. For a neon-filled counter the maximum permissible voltage between any two electrodes (excluding the anode) is accordingly 140 V. With double rectangular pulse drives the pulse amplitudes must clearly be substantially less than this. Consideration of Fig. 7.4 (*b*) will show, however, that with integrated pulse drive the maximum pulse amplitude can exceed 140 V by a

few volts because with finite rise and fall times on the Guide B pulse the full pulse voltage is never developed between Guides A and B.

Spurious Discharges Due to Sharp Pulses

Very rapid rise and fall times are to be avoided. They can cause the anode–cathode voltage to change so rapidly that the discharge cannot establish itself sufficiently quickly in the appropriate gap. As a result, the anode–cathode voltage across this gap rises above the usual 190 V. This temporary increase in voltage may be sufficient to cause breakdown either to the wrong cathode (i.e. stepping in the reverse direction) or between the leads under the top ceramic of the tube.

A somewhat similar phenomenon occurs if the anode load resistor is not connected directly to the tube base. Capacitance between anode load and ground may excessively restrict the rate of rise of anode potential required to follow a rising guide potential. The voltage across the gap will then fall below 190 V and the discharge will extinguish. Subsequently the anode voltage will rise towards the h.t. rail until breakdown occurs – not necessarily in the required gap. These effects are avoided by restricting the rise and fall rates of guide pulses to less than 100 V/μsec and the rise of h.t. to a time-constant of 1 msec. This latter restriction is implicit in most h.t. supplies. If a switch is used in the h.t. supply, however, decoupling should be provided by 10 kΩ and 0·1 μF. The reset circuit should also have a time-constant of at least 1 msec.

Output Pulses

An output pulse may be obtained by connecting a resistor R_K between any index cathode and ground. When the discharge rests on this cathode its potential rises due to the flow of cathode current through this resistor. If the anode current is not to be shared between the cathode and an adjacent guide it is necessary to restrict the value of R_K so that the cathode potential remains at least 10 V below the guide bias. With a bias of +40 V, therefore, stepping may become uncertain if the cathode potential is allowed to exceed +30 V. If a larger output pulse is required this may be obtained by returning the cathode resistor to a negative potential not exceeding −20 V. This technique is not applicable to adjacent index cathodes, and is therefore not generally used with selector tubes. To accommodate tube ageing, the negative bias is better restricted to −12 to −15 V. Use of a negative cathode bias requires a corresponding increase in guide pulse amplitude and/or reduction in guide bias.

Manufacturers differ slightly in their recommendations regarding the permissible limits for output cathode bias and pulse amplitude. This is because, apart from differences between competitive tubes, the whole question is one of compromise between performance, component tolerances, and tube ageing.

Mullard Ltd. [5] have reported that the limitation on maximum

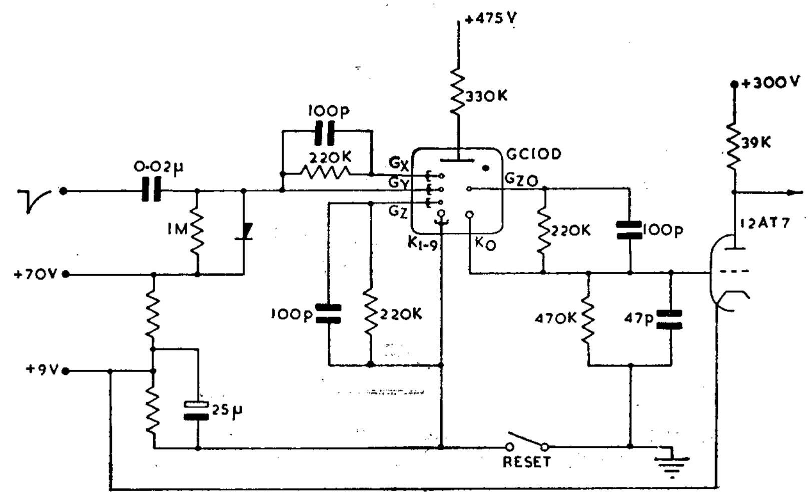


Fig. 7.6. Guide bias, drive, and interstage coupling for single-pulse 'Dekatron'.

stepping speed of a tube is normally imposed by the necessity to step to cathodes at different potentials. If there is no necessity to obtain output pulses the index cathodes may all be returned to ground, and a much higher rate of stepping can then be achieved. It was found that in this condition all samples of the Z504S would operate reliably up to 18 kc/s, whereas normally they are not recommended for use above 5 kc/s. As a result of this observation, the circuit of Fig. 7.7 was developed. Use is made of the fact that the stepping tube is transferring a *current* from one cathode to another. The current received by the output cathode is applied to the base of a transistor, and a 12-volt pulse is then delivered from the collector. This happens to be of the opposite polarity from that produced by the conventional cathode-resistor output. A standard pulse-differentiating inter-stage coupling may still be used, however, if the transistor is connected to K_9 instead of K_0. The transistor then delivers the required positive step as the glow leaves K_9 instead of when it alights on K_0.

Due to effects discussed below, the maximum speed of this circuit is reduced after the tube has been operating in a static condition for some time. Even then, however, a stepping speed of 10 kc/s can be obtained without any reduction of tube life.

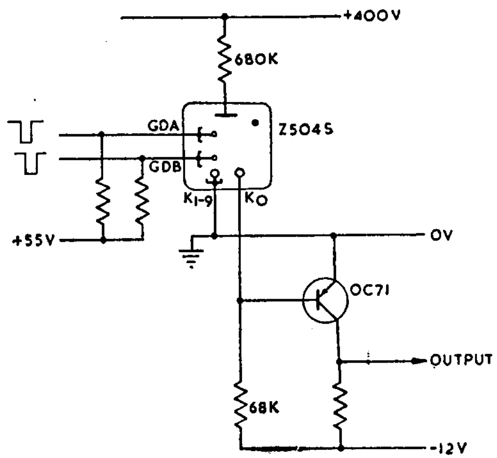


Fig. 7.7. Interstage coupling operated by cathode current rather than cathode potential. A higher maximum stepping speed results.

Tube Life

It was noted earlier that the glow transfer time of a new neon-filled tube may be about 25 μsec, but that this will increase to a stable value of about 75 μsec. This change may occur during the first 1,000 hours of tube life. With correct use, the glow transfer time may thereafter remain practically constant for many thousands of hours. Eventually, however, the transfer time begins to rise again. When it has increased beyond the value specified by the manufacturer the life of the tube is technically at an end, even though stepping may still be effected by using guide pulses of longer duration.

If the glow rests continuously on a single cathode the tube life may be only 2,000–3,000 hours. Since tube life depends largely on cathode condition, it is not surprising that a tube with 30 cathodes should show a life of up to 90,000 hours provided it is cycled continuously with equal dwell times on each cathode. In practice, tube life is a much more complicated function of operating conditions.

When the glow rests on a particular cathode nickel is sputtered from that cathode on to adjacent guide electrodes. This sputtered-on nickel

has properties which differ from those of the pure nickel. If the guides receive an excess of sputtered nickel, therefore, the tube will refuse to step with guide pulses of normal dimensions. Excessive anode current will clearly accelerate the sputtering process. The minimum permissible tube current is determined either by inadequate priming of the adjacent anode-guide gaps or by the danger of entering an unstable negative-resistance part of the discharge characteristic. The recommended anode current exceeds these danger levels as little as possible so that sputtering is reduced to a minimum.

The choice of guide bias also has an important effect on contamination of the guides by sputtered nickel. A relatively low guide bias voltage will allow 5 or 10% of the anode current to flow to either guide adjacent to the active cathode. If the guide bias is made high – in order to obtain large output cathode pulses, for example – the anode current is no longer shared in this way. Since this small current helps to restrict contamination of the guides, it is desirable to restrict the guide bias to a moderate value. This is particularly important in a scaler counting random pulses, since in this application the glow necessarily rests on the index cathodes for a much greater proportion of the time than on the guides. Counters working on a constant frequency should, if possible, be designed so that the discharge rests for equal proportions of the cycle on cathode, Guide A and Guide B. This principle should be observed in each tube. In all applications a bias of about +40 V will give maximum tube life.

After it has been prematurely aged by prolonged static operation a neon-filled tube may be rejuvenated very largely by continuous rapid stepping for an hour. This may require abnormally long guide pulses at first, but the transfer time will be reduced as the contaminated guides are sputtered clean. This recovery is not observed in tubes filled with gas mixtures other than neon. A decade counter using several similar neon-filled tubes lends itself to a simple preventive maintenance technique: once a month remove the fastest-stepping tube, move each of the others up one decade so that each is operating ten times faster than before, finally replace the tube first removed so that it now operates in the slowest-stepping position.

A tube which is not being used should either be switched off (by removing the anode supply) or cycled continuously (e.g. at power-supply frequency).

Storage at above 60° C leads to contamination of the electrodes due to out-gassing of the glass. If the tube is operating continuously, temperatures up to 100° C have no ill effect. Above 60° C it is therefore

better to keep a tube cycling continuously rather than completely inert.

Summarizing:

(1) Use a guide bias of +40 V whenever possible.
(2) Operate at the recommended anode current.
(3) Keep the tube cycling.
(4) For each tube, make the guide-pulse durations equal the cathode dwell time at the maximum speed at which that tube has to operate.
(5) In a decade scaler interchange similar tubes regularly.
(6) Do not store at high temperatures.

Guide Bias Arrangements

If the simple circuit of Fig. 7.8 (*a*) is used to provide the positive guide

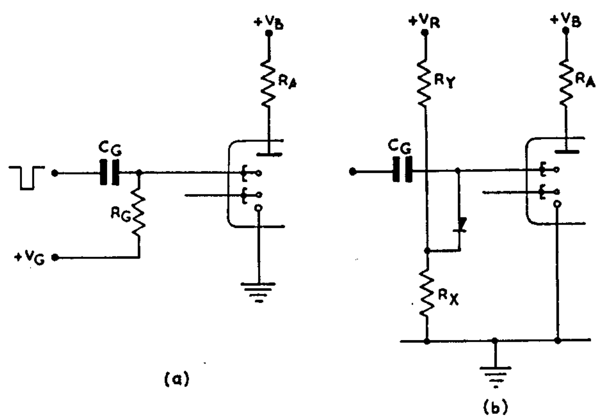


Fig. 7.8. The guide bias arrangement shown at (*a*) leads to autobiasing problems which are solved in the arrangement at (*b*).

bias, V_G, difficulties are soon encountered. A train of rectangular pulses applied through C_G would, even with the tube non-conducting, lead to a redistribution of potentials so that the *mean* guide potential assumes the value V_G, the guide swinging above as well as below this potential. The limits of the excursions of guide potential would therefore depend not only on the bias V_G and the amplitude of the applied pulses but also on the width and the recurrence rate of the pulses.

When the tube is conducting a further complication arises. While the drive is holding the guide negative so that the glow rests on the guide, current flowing from anode to guide charges the capacitor C_G. To allow this charge to leak away, C_GR_G must be made small compared with the interval between pulses. If this is not done the rectifying action causes

the guide to bias itself almost to cut-off so that stepping does not occur. On the other hand, a small value of C_GR_G causes the guide pulse to be differentiated so that the guide may not be held negative sufficiently long to effect stepping.

All these difficulties are best overcome by using a diode to clamp the guide to a suitable bias potential as shown in Fig. 7.8 (*b*). When a negative pulse is applied via C_G, the diode is non-conducting. The duration of the negative pulse is thus limited only by the charging of C_G by anode current flowing into the guide. After a time, t_P, C_G charges so much that the pulsed guide is no longer the most negative electrode. The anode current is then shared with an adjacent electrode. From such considerations the minimum value of C_G for a given value of t_P is given by:

$$C_{G(\min)} = I_A t_P / V_{GT(\text{peak})} \tag{7.1}$$

where $V_{GT(\text{peak})}$ = peak guide transfer voltage = $(V_P - V_G)$ and C_G is in pF, I_A in μA and t_P in μsec.

With this value of C_G, however, V_{GT} has fallen to zero at the end of the time t_P. A faster and more positive transfer is therefore obtained by making C_G much larger so that V_{GT} remains almost unchanged throughout the pulse. In a double-rectangular pulse drive, therefore, it is common to make C_{GA} nearly two orders larger than $C_{G(\min)}$,

i.e. $$C_{GA} \approx I_A t_{PA} \tag{7.2}$$

This ensures that the discharge transfers quickly from cathode to Guide A before the Guide B pulse commences. This done, the return of Guide A to the positive bias helps to transfer the discharge to Guide B. C_{GB} may safely be made considerably smaller than C_{GA}. Since positive bias prevents Guide A from sharing anode current with Guide B, charging of C_{GB} assists transfer to the next index cathode.

Whether or not a diode is used, it is important to arrange that the time-constant C_GR_G (in Fig. 7.8 (*a*)) or $C_G[R_XR_Y/(R_X + R_Y)]$ (in Fig. 7.8 (*b*)) is less than the interval between successive pulses. Unless this is done, C_G will still carry a substantial charge at the start of the next pulse.

Fig. 7.9 illustrates the application of these principles to the biasing of a GS10D for double-rectangular pulse drive.

Resetting

Resetting to a partcular index cathode is effected simply by arranging that its potential is 120 V below any other electrode. The voltage between

N

anode and the chosen cathode then exceeds the breakdown voltage and transfer occurs within 5 μsec.

This may be done in two ways:

(1) The chosen cathode may receive a −120 V pulse of at least 5 μsec duration.

(2) All other cathodes and guides may temporarily be raised 120 V above the chosen cathode.

In a decade counter the first method leads to generation of 'carry' pulses during resetting to zero. Instead of the desired 0000 the indicated

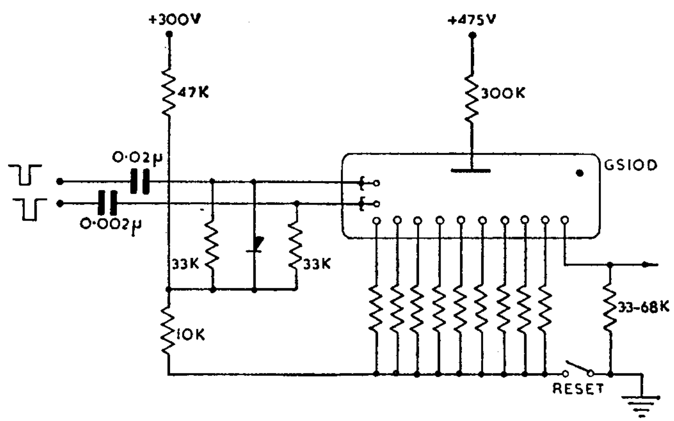


Fig. 7.9. Guide bias arrangements for double-rectangular pulse drive of GS10D.

count may thus become 1110. This arises as follows: the negative pulse is applied to all K_0 cathodes and the discharge in each tube accordingly moves to this electrode. As this is the output cathode used for interstage coupling, however, the decay of the reset pulse allows each stage to pass a 'carry' pulse into the next.

One way of overcoming this limitation is found in resetting to 9990. The carry pulse from the units then passes through the instrument and changes the display to 0000. This means that, except for the 'units' stage, tubes must be used having K_0 and K_9 separate from K_{1-8}.

An alternative solution due to Whelan [6] comprises using interstage couplings in which the 'carry' pulses are due to the fall of potential of K_9 rather than the rise of potential of K_0.

The second method of resetting does not give rise to the same problem. When all guides and cathodes other than K_0 are raised to +120 V the discharge in each tube moves to K_0 and produces a 'carry' pulse.

The succeeding stage is unable to respond to the carry pulse, however, so long as the reset condition persists. When the carry pulses have decayed the guides and cathodes may return to their normal potentials without a change of count taking place.

Input Drives

For the first stepping tube of a counting chain, a drive circuit is required providing guide pulses and bias of type and dimensions satisfying the considerations outlined above. The most appropriate circuit will depend on the source and nature of the signal, but in most cases signal shaping must be effected by valves or transistors. Tube manufacturers offer to provide advice on applications not already covered by their published literature [7, 8, 9]. General principles may be summarized, however, according to the type of signal to be counted:

(a) Continuous Sine Wave

This is the simplest case. Provided the input frequency does not vary greatly and is applied continuously, the simple circuit of Fig. 7.10 may

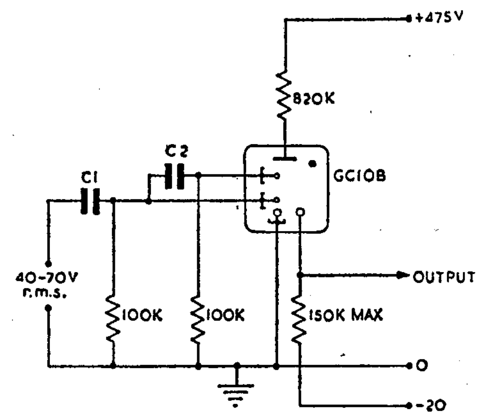


Fig. 7.10. Input drive for operation of GC10B from continuous sine wave.

be used. Here the guide bias is provided by the self-rectifying action of the guides. The correct bias is thus established only after the tube has received several consecutive pulses at the frequency, f, for which C_1 and C_2 have been chosen.

$$C_1 \approx 2C_2 \tag{7.3}$$

$$C_2 \approx (5/f) \text{ microfarads} \tag{7.4}$$

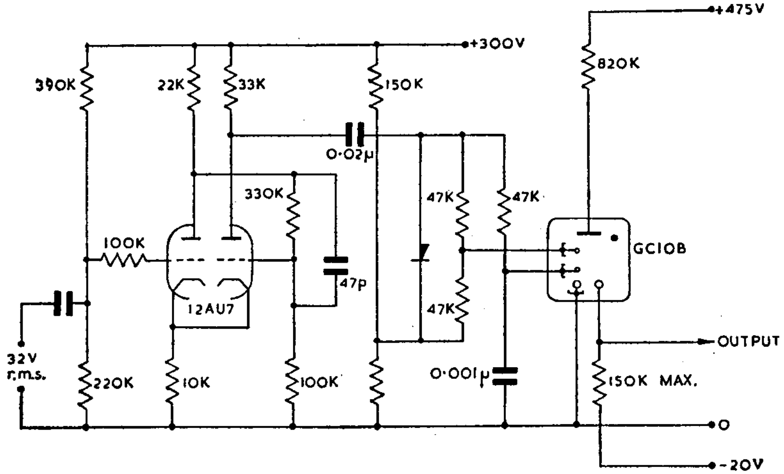


Fig. 7.11. Reshaping circuit converting train of sine waves to rectangular pulses for counting by GC10B.

(b) Sine-wave Train

When it is required to count the cycles of a train of sine wave, counting of the initial cycles may be assured by reshaping the wave to a rectangular form and then using an integrated pulse drive as in Fig. 7.4 (*b*). A suitable drive circuit is shown in Fig. 7.11.

(c) Step Function

When the input is in the form of a step function, or a rectangular pulse of unspecified duration, it is not possible to choose a suitable time-constant for an integrated pulse drive. The most convenient drive arrangement is then usually the combination of differentiated and integrated pulses (Fig. 7.4 (*c*)). Inter-stage couplings (p. 190) are usually of this type. In appearance the circuit may strongly resemble Fig. 7.11. However, whereas in Fig. 7.11 the trailing edge of the Guide A pulse is defined by cut-off of the Schmitt trigger, a circuit operating from a step function uses differentiation in the coupling capacitor to provide a trailing edge to the Guide A pulse.

(d) Pulse

Sharp pulses may be converted to well-defined rectangular form and registered by an integrated-pulse drive, as in Fig. 7.12. This circuit will operate from a negative-going pulse of 30 V or more and having a rate of fall (on the leading edge) of at least 100 V/μsec.

Slower pulses may be shaped by the circuit of Fig. 7.11.

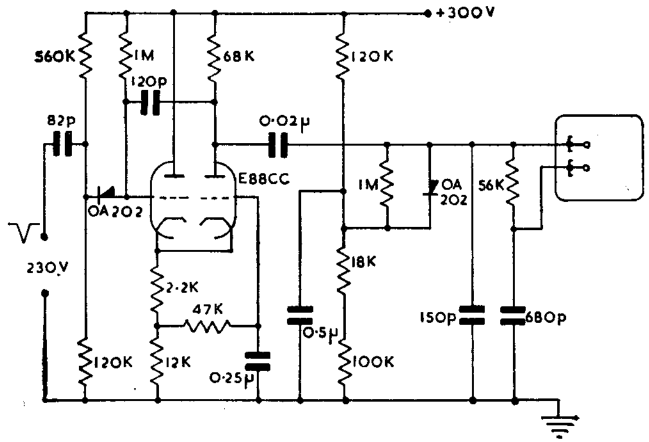


Fig. 7.12. Drive circuit converting sharp pulses to rectangular form which then provides integrated-pulse drive to stepping tube at right.

(e) Random Pulse

Nuclear scalers are required to register pulses having a random rate of arrival. An important consideration is the resolving time of the counter. Consecutive pulses arriving with a smaller time separation will be registered as a single count. Since gas-filled stepping tubes have a somewhat restricted resolving time, the resulting statistical error can be significant.

For example, the Z505S can count regular pulses at up to 50 kp/s. But because it resolves input pulses only if they are 20 μsec apart or more, randomly spaced pulses must be counted at a much lower mean rate if reasonable accuracy is required. If 1 in N of the input pulses arrive at a closer spacing, the indicated count will be low by $100/N$ per cent. The performance can be greatly improved by adding a binary stage before the input. This serves three purposes:

(i) It defines the drive pulse dimensions.

(ii) It reduces the mean rate of stepping by a factor of 2. Thus a *regular* pulse train of 100 kp/s may be counted by a Z505S preceded by a binary stage.

(iii) If the probability of arrival of two consecutive pulses within 20 μsec is 1 in N, the probability of arrival of three pulses in the same time is 1 in N^2. For a given statistical error of 1 in N, the addition of a suitable binary stage therefore allows an increase of mean counting rate by a factor N.

The design of fast binary counters has been discussed at length elsewhere [10, 11].

Inter-stage Couplings

The output cathode of a tube provides a positive step function when the glow alights on it. This step must be amplified and shaped before it can be used to drive the next stepping tube in a decade scaler. If the first tube is fed with random pulses it is not possible to specify the time

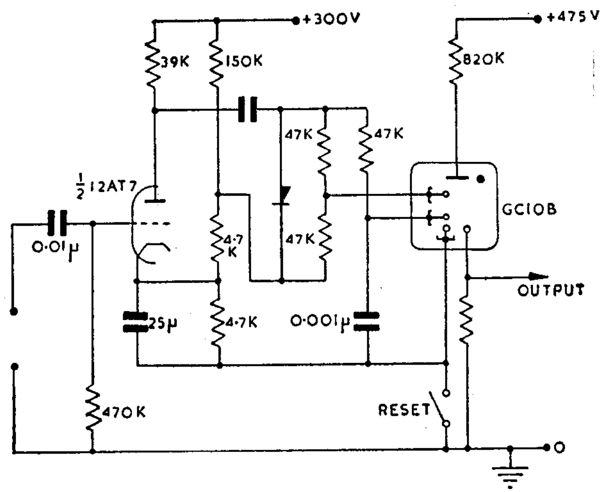


Fig. 7.13. Interstage coupling suitable for use between two neon-filled stepping tubes.

for which the glow will dwell on the output cathode. Thus, although the output cathode delivers a rectangular pulse, it is not practicable merely to invert and amplify this pulse to drive Guide A of the next stage and to integrate the rectangular pulse to drive Guide B. As explained at *Input Drives* above, the step function calls for a differentiated-plus-integrated pulse drive circuit.

Even when the input is at a constant frequency, pulse shaping may be necessary. Suppose, for example, a GC10D is driven at its maximum stepping rate of 20 kp/s. The glow then rests on the output cathode for about 20 μsec every 500 μsec. The duration of the 20 μsec output pulse is too short to operate a neon-filled stepping tube, such as a GC10B, even though the tube in this succeeding stage is not required to count at above 2 kp/s. Although the 20 μsec pulse could be used to operate a

second GC10D, it is more usual to stretch the 'carry' pulse from the first tube so that it lasts for the 80 μsec required by the slower tube.

The inter-stage coupling shown in Fig. 7.6 shapes the pulses in the following way. When the glow alights on the output guide, G_{Z0}, the flow of anode current through the 470 kΩ resistor causes the grid potential of the 12AT7 to rise. Grid current flows and clamps K_0 to

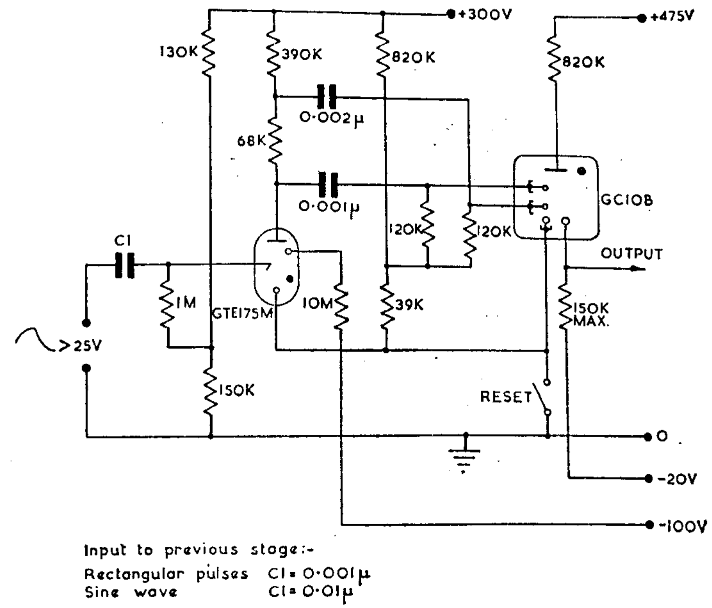


FIG. 7.14. Interstage coupling using trigger tube. With capacitor values indicated, it may also be used as an input stage.

approximately the +9 V bias to which the 12AT7 cathode is returned. When the glow leaves K_0 grid current ceases to flow, but the potential of the 12AT7 grid is temporarily maintained by the 47 pF (and Miller capacitance) of the grid. Thus, although the 12AT7 anode potential falls abruptly immediately the glow reaches G_{Z0}, it does not recover until several tens of microseconds after the glow has left K_0. The 12AT7 anode is connected *via* 0·02 μF to a guide bias and integrating circuits identical to those of Fig. 7.11.

For coupling between two neon-filled stepping tubes, no pulse stretching is required. The interstage coupling must still differentiate the pulse, however, and in the circuit of Fig. 7.13 this is done by the coupling capacitors.

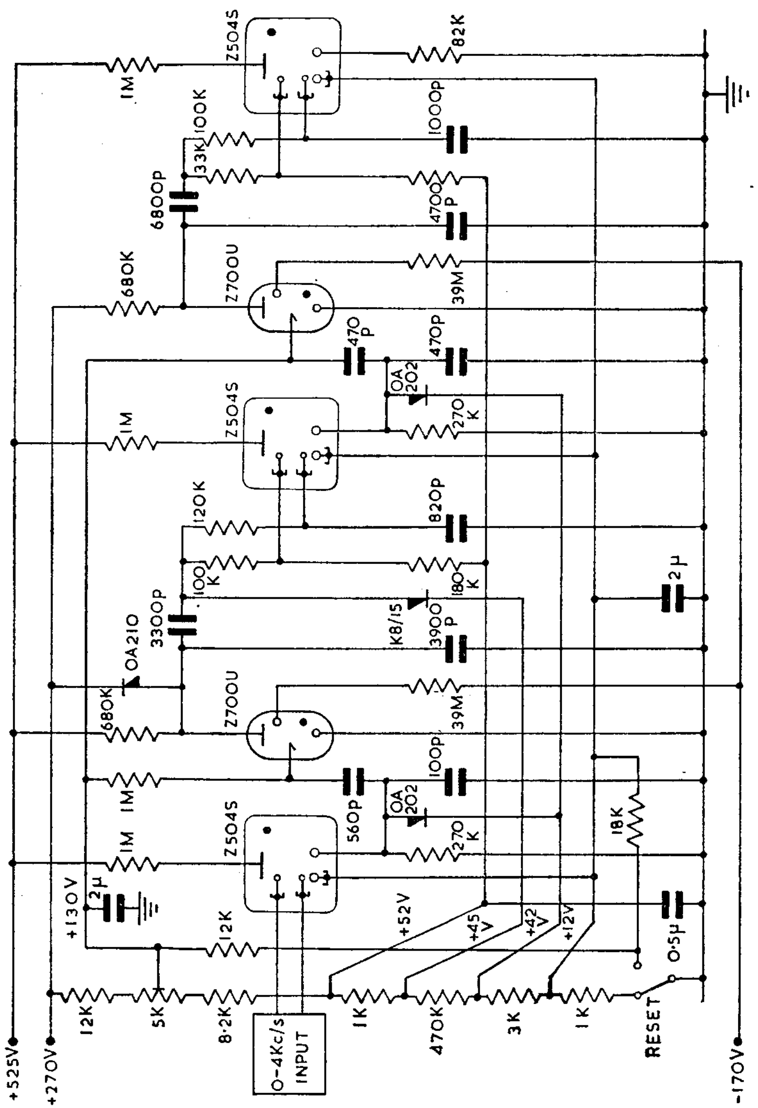


Fig. 7.15. Mullard circuit for decade scaler using trigger tubes for inter-stage coupling.

Inter-tube coupling may be effected by trigger tubes. Fig. 7.14 shows a circuit capable of handling up to 500 'carries' per second. In Fig. 7.14, the trigger and anode circuits each operate in the self-quenching mode. When the trigger tube fires, Guide A is depressed to a lower potential than Guide B. Consequently, the discharge moves in a clockwise direc-

tion. The stepping tube anode current flows into Guide A and recharges C_2, so that after about 80 μsec Guide A has risen to ground potential. Transfer then occurs to Guide B, which is still some 60 V negative. C_3 recharges in the same way and, when both Guide A and Guide B are positive, transfer occurs to the next index cathode.

Fig. 7.15 shows a Mullard circuit in which clamping diodes are used

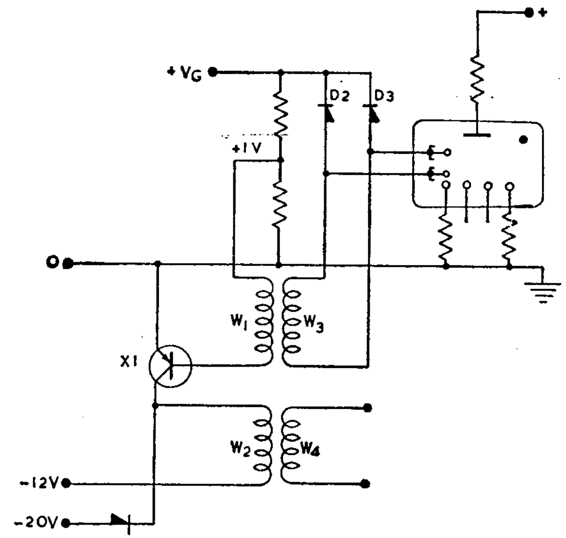
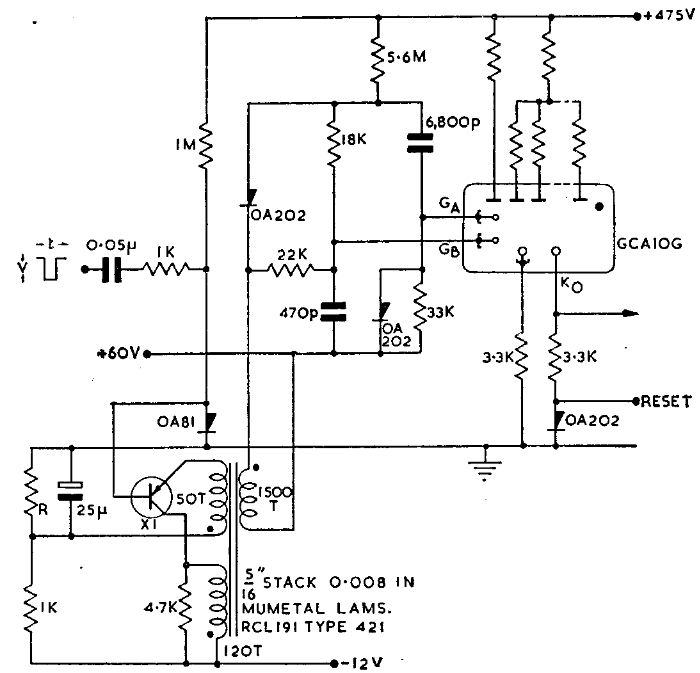


Fig. 7.16. Transistor blocking oscillator producing double-rectangular pulse drive for stepping tube.

to allow the Z700U to serve as an interstage coupling amplifier handling up to 400 'carries' per second. The trigger tube provides a negative-going pulse which is differentiated and integrated in a conventional manner for application to the stepping tube guides. Although this circuit calls for a rather large number of different bias supplies, they are readily obtained from a single bleeder chain.

Transistors may also be used for driving and coupling stepping tubes [12, 13, 35, 36, 37]. Warman and Bibb [12] have described drive and reset units in which the relatively large pulses are obtained from transistor blocking oscillators. They operate several stepping tubes from a transistor d.c. converter delivering 500 V at 5 mA. Even in circuits provided only with low-voltage supplies for transistors, they are thus able to take advantage of the circuit economy and simple read-out offered by the cold cathode stepping tube.

The blocking oscillator drive used by Warman and Bibb is shown in Fig. 7.16. The base of transistor X_1 is normally biased to about $+1$ V, so that continuous oscillation will be inhibited. When an input pulse is applied to winding W_4 regeneration from the collector winding W_2 to base winding W_1 causes the transistor to switch on and 'bottom'. The collector remains at ground potential until the transformer core



X_1	R	t_{MIN}	V_{MIN}	f_{MAX}
OC77	27Ω	10μs	5V	2Kc/s
OC83	12Ω	20μs	10V	5Kc/s

Fig. 7.17. Transistor blocking oscillator providing differentiated and integrated pulses to Guides A and B respectively.

saturates. The regenerative action then reverses and the collector potential swings negative until caught at -20 V by diode D_1. As the transformer back e.m.f. collapses, the collector returns to -12 V. When an input is applied to W_4, therefore, the collector potential swings from -12 V to ground and thence to -20 V before returning to -12 V. As a result, the winding W_3 delivers a pulse of about 100 V, first of one polarity and then of the other. Diodes D_2 and D_3 clamp the stepping-

tube guides to the bias potential V_G whenever they tend to become positive. Thus D_2 conducts while GDA is pulsed negatively and then, as the output of W_3 reverses polarity, D_3 conducts while GDB is pulsed negatively. The stepping tube therefore receives double rectangular guide pulses and can operate at its maximum speed. Warman and Bibb give full details of the transformers used in the blocking oscillator and the d.c. converter.

An Ericsson circuit (Fig. 7.17) uses a blocking oscillator to drive a GCA10G or GSA10G. This circuit requires only one negative supply. A single pulse from the transformer is differentiated for application to Guide A and integrated for application to Guide B.

Sadowski and Cassidy [36] describe a rather simpler arrangement, but do not specify for which type of tube it is designed.

Birk and others [37] describe a scaler using high-speed stepping tubes with inter-stage couplings provided by high voltage switching transistors. Only the first stage of their scaler is driven by a blocking oscillator. In order to achieve a high switching rate, reflected capacitive loading is minimized by using a high-voltage transistor and a transformer of small turns ratio. Moreover, the effective guide input capacitance of the stepping tube ($\approx$36 pF) is reduced by returning cathodes 1–9 to the base of the transistor rather than to ground. This places the guide–cathode capacitance in a positive feedback path so that it accelerates rather than retards switching. By these techniques the EZ10B in the first stage is driven at up to 500 kc/s.

The Ericsson GS10J has been developed specifically for transistor drive. It requires a guide pulse amplitude of only 24 V, which is readily provided by a transistor flip-flop. Thus, although a second transistor is required, no transformer is needed.

Reversible Counters

When a reversible scaler is required an interstage coupling is necessary capable of delivering rectangular drive pulses in correct sequence to the following stage. For this purpose a special reversible stepping tube has been devised [14]. In most respects it resembles a normal double-pulse tube, but separate connexions are made available to the two guide electrodes situated between cathode K_9 and K_0. These 'routing guides', RG_A and RG_B, are returned to Guides A and B through 22 kΩ resistors, as shown in Fig. 7.18. When the tube steps from K_9 to K_0 the discharge briefly rests on RG_A and RG_B and a potential appears across either resistor in turn. A direct-coupled differential amplifier inverts and amplifies the RG_A pulse and applies it to Guide A of the next stepping

tube. A second amplifier transfers the RG_B pulse to Guide B of the following stage. If the first tube is being driven in the 'add' direction, therefore, staggered double rectangular pulses in the 'add' sequence are passed to the following stage as the glow moves from K_9 to K_0. When the first tube is driven in reverse the sequence of 'carry' pulses reverses

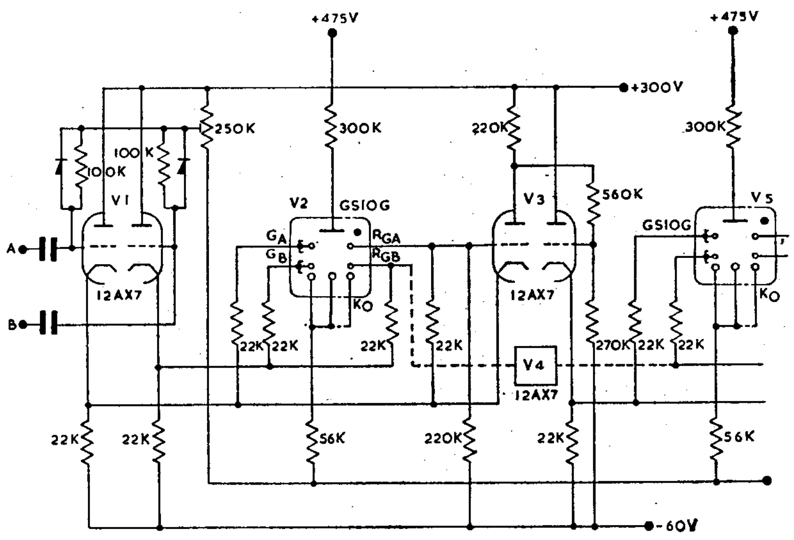


Fig. 7.18. Reversible decade scaler in which routing guides are used to determine sequence of 'carry' pulses applied to subsequent stage.

also, i.e. a 'subtract' signal is sent to the second tube each time the first moves from K_0 to K_9.

The system of inter-stage coupling may be extended to a counter of several decades.

A circuit due to Oxley [15] provides a reversible decade scaler using the simpler tubes in which no separate connexions are provided to routing guides. Oxley's circuit uses two univibrators, one providing 'add' pulses, the other 'subtract' pulses. When an 'add' pulse causes the glow to move on to the K_0 cathode in one stage, coincidence of pulse and rise of K_0 potential trips the 'add' univibrator of the next stage. The 'subtract' univibrator is tripped by coincidence of a 'subtract' pulse and the fall of K_0 potential due to glow leaving K_0. The circuit is scarcely more complex than that needed for the special reversible tubes, and Oxley

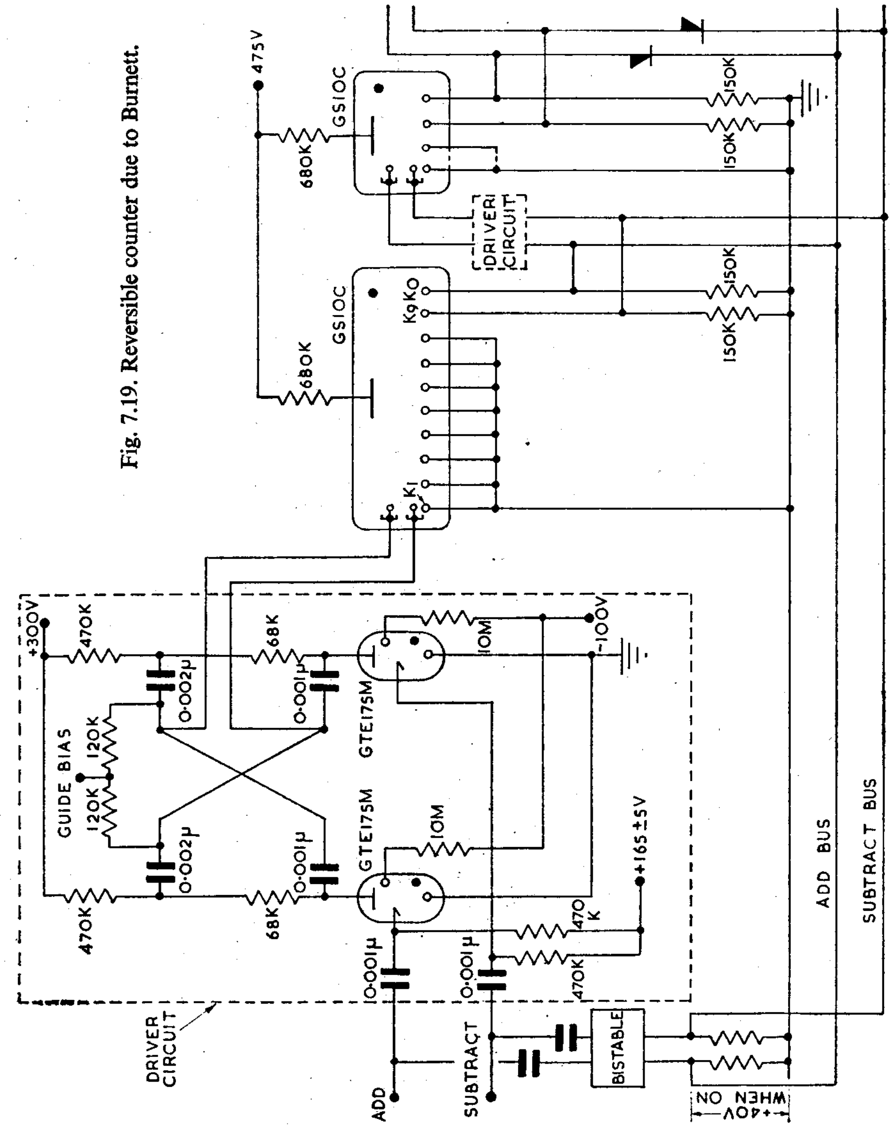


Fig. 7.19. Reversible counter due to Burnett.

claims it need not be subject to false counts due to rapid reversals of direction of counting.

A simple circuit arrangement described by Burnett [16] may be used when the 'add' and 'subtract' inputs are separate, provided their separation in time exceeds both the normal resolution time and the time re-

quired for a 'carry' to pass right through the counter chain, from the 'units' to the most significant decade. Burnett's circuit is shown in Fig. 7.19. The first pulse of either input sets a bistable to an 'add' or 'subtract' condition, so that diode gates pass only the appropriate 'carry' pulses from one tube to the driver circuit of the next. Each driver stage comprises two trigger-tube circuits comparable with that of Fig. 7.14. One tube produces an 'add' drive from the preceding K_0 output, the other a 'subtract' drive from the K_9 output. Since the bistable is reset on the leading edge of the first pulse arriving with a reversal of direction, the diodes are always biased in the correct way before the glow transfers to K_9 or K_0. The counter is thus able to handle rapid reversals of count (e.g. 8, 9, 0, 9, 8) subject to the limitations specified above.

Anoded-Cathode Circuit Design

A typical stepping tube may operate with an anode current anywhere between 250 and 550 μA. The lower limit of the range is set by instability of the discharge, the upper by reduction of cathode life due to sputtering. For given values of anode and cathode resistance, R_A and R_K, the anode current I_A is given by:

$$I_A \approx \frac{V_B + V_K - V_M}{R_A + R_K} \tag{7.5}$$

where V_K is the negative bias to which R_K is returned.

From this, it might appear that quite wide tolerances can be accepted on V_B and/or $(R_A + R_K)$. The following analysis shows that a number of factors influence circuit design, and it is only with care that useful tolerances may be obtained on all parameters.

Since the output from a stepping tube is due to the passage of I_A through R_K, a design based on the full working range of I_A would lead to an output pulse varying more than 2 to 1. Where the output pulse must be of a defined amplitude, a catching diode should be used. Fig. 7.20 shows a convenient arrangement in which the grid–cathode gap of an inter-stage coupling amplifier acts as a diode catching K_0 when it reaches ground potential.

It will be seen shortly that in practice the output cathode current may not vary over the full range of I_A. For some applications the output pulse may be sufficiently well defined by using rather close tolerances on V_B, R_A, and R_K.

The highest value of anode current is drawn when the glow rests on a negatively pulsed guide. Knowing the positive bias, V_G, on the guides and the amplitude, V_P, of the negative-going guide pulse, one can write:

$$I_{A(\max)} \geqslant \frac{(1 + r)V_B - V_M - V_G + V_P}{(1 - w)R_A} \tag{7.6}$$

where r and w are acceptable fractional tolerances on V_B and R_A respectively.

The minimum value of anode current occurs when the glow rests on

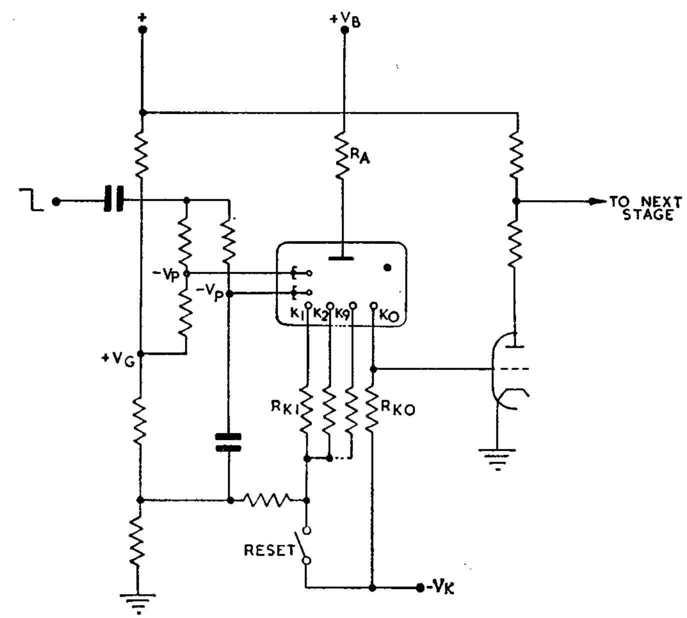


Fig. 7.20. Typical stepping tube circuit on which design procedure is based.

the cathode reaching the most positive potential. If this potential is V_0 above ground, then:

$$I_{A(\min)} \leqslant \frac{(1 - r)V_B - V_M + V_0}{(1 + w)R_A} \tag{7.7}$$

The lowest acceptable value of nominal anode supply voltage is obtained by combining Relations (7.6) and (7.7).

$$V_{B(\text{nom})\,(\min)} = \frac{(1 - w)I_{A(\max)}(V_M + V_0) - (1 + w)I_{A(\min)}(V_M + V_G - V_P)}{(1 - r)\,(1 - w)I_{A(\max)} - (1 + r)\,(1 + w)I_{A(\min)}} \tag{7.8}$$

and the corresponding value of R_A is then

$$R_A = \frac{(1 + r)V_{B(\text{nom})\,(\min)} - V_M - V_G + V_P}{(1 - w)I_{A(\max)}} \tag{7.6a}$$

In practice, it is usual to increase R_A to the next preferred value. This will increase the value of V_B needed, but allow some latitude in choice of the precise value of V_B. So far as possible a nominal value of V_B should be chosen which is midway between the limiting values corresponding to Relations (7.6) and (7.7).

The output voltage, $(V_K + V_0)$, developed across a cathode load resistor, R_K, is given by:

$$(V_K + V_0) = I_K R_K$$
$$\approx 0{\cdot}85 I_A R_K \qquad (7.9)$$

There is, however, a limit to the maximum voltage, V_0, to which an output cathode can rise. If R_K is increased until the cathode approaches the guide bias potential, then an increasing proportion of the anode current flows to the adjacent guides. There is thus practically no increase in V_0 if R_K is increased beyond this value. Moreover, because of the current-sharing, the tube may not step correctly.

If V_0 is the critical value of output cathode voltage (measured above ground) the corresponding value of R_K may be expressed generally as:

$$R_{K(crit)} = \frac{(V_K + V_0)}{(V_B - V_M - V_0)} \cdot \frac{R_A}{0{\cdot}85} \qquad (7.10)$$

When no catching diode is used, V_0 should not exceed $(V_G - 10)$. Then:

$$R_{K(max)} = \frac{(V_K + V_G - 10)}{[(1+r)V_B - V_M - V_G + 10]} \cdot \frac{(1-w)}{(1+w)} \cdot \frac{R_A}{0{\cdot}85} \qquad (7.10a)$$

If a catching diode is used to restrict the value of V_0 to less than $(V_G - 10)$ the value of R_K should be made sufficiently large to ensure correct limiting action. Then:

$$R_{K(min)} = \frac{(V_K + V_0)}{[(1-r)V_B - V_M - V_0]} \cdot \frac{(1+w)}{(1-w)} \cdot \frac{R_A}{0{\cdot}85} \qquad (7.10b)$$

Typically, R_K might be made double this value so that the tolerance on R_K could be much more than w.

Design Procedure for Anode–Cathode Circuit of Stepping Tube

(*a*) From the manufacturer's data, set out:

Maximum permissible anode current, $I_{A(max)}$
Minimum permissible anode current, $I_{A(min)}$
Guide bias voltage, V_G
Maximum amplitude of negative guide pulse, V_P

Anode–cathode voltage drop, V_M
Minimum anode supply voltage, $V_{B(min)}$

(*b*) Set out the following data for circuit considered:

Maximum positive voltage of output cathode, V_0
Fractional tolerance, r, required on anode supply voltage, V_B
Fractional tolerance, w, required on anode load resistor, R_A

(*c*) From Equation (7.8), determine the minimum nominal supply voltage, $V_{B(nom)\,(min)}$, which will satisfy the above requirements.

$$V_{B(nom)\,(min)} = \frac{(1-w)I_{A(max)}(V_M + V_0) - (1+w)I_{A(min)}(V_M + V_G - V_P)}{(1-r)(1-w)I_{A(max)} - (1+r)(1+w)I_{A(min)}} \qquad (7.8)$$

(*d*) If $V_{B(nom)\,(min)}$ is unacceptably high, use a smaller value of r and/or w and repeat operation (*c*).

(*e*) Check that

$$V_{B(nom)\,(min)} \geqslant V_{B(min)}/(1-r) \qquad (7.11)$$

If necessary, increase $V_{B(nom)\,(min)}$ to satisfy Relation (7.11).

(*f*) Calculate the value of R_A corresponding to $V_{B(nom)\,(min)}$.

$$R_A = \frac{(1+r)V_{B(nom)\,(min)} - V_M - V_G + V_P}{(1-w)I_{A(max)}} \qquad (7.6a)$$

(*g*) If R_A does not happen to be a preferred resistance value, choose the nearest preferred value, R'_A above the calculated value.

(*h*) For the selected preferred value of R'_A, calculate the limiting values of $V'_{B(nom)}$.

$$V'_{B(nom)\,(min)} = \frac{1}{(1-r)}\{(1+w)R'_A I_{A(min)} + V_M - V_0\} \qquad (7.7a)$$

$$V'_{B(nom)\,(max)} = \frac{1}{(1+r)}\{(1-w)R'_A I_{A(max)} + V_M + V_G - V_P\} \qquad (7.6b)$$

(*i*) Choose V_B between these two limits and as nearly as possible midway between them.

$$V_B \approx \tfrac{1}{2}(V'_{B(nom)\,(min)} + V'_{B(nom)\,(max)}) \qquad (7.12)$$

(*k*) Where a catching diode is used to limit the output cathode to V_0, choose $R_K > R_{K(min)}$ given by Equation (7.10b).

$$R_{K(min)} = \frac{(V_K + V_0)}{[(1-r)V_B - V_M - V_0]} \cdot \frac{(1+w)}{(1-w)} \cdot \frac{R_A}{0{\cdot}85} \qquad (7.10b)$$

(By putting $R_K \approx 2R_{K(min)}$, the tolerance on R_K may be increased.)

(*l*) Where no catching diode is used, choose $R_K \leqslant R_{K(max)}$ given by Equation (7.10a).

$$R_{K(max)} = \frac{(V_K + V_G - 10)}{[(1+r)V_B - V_M - V_G + 10]} \cdot \frac{(1-w)}{(1+w)} \cdot \frac{R_A}{0{\cdot}85} \quad (7.10a)$$

EXAMPLE 7.1 Anode–cathode Circuit of Stepping Tube

A Z504S stepping tube is required to operate in the circuit of Fig. 7.20. Determine the value of anode supply voltage for a tolerance of $\pm 12\%$. Also determine values and tolerances of anode and cathode resistors.

(*a*) From the manufacturer's data:

$I_{A(max)} = 0{\cdot}525$ mA $\quad V_G = 40$ V $\quad V_K = 12$ V $\quad V_{B(min)} = 375$ V
$I_{A(min)} = 0{\cdot}250$ mA $\quad V_P = 100$ V $\quad V_M = 195$ V

(*b*) For circuit of Fig. 7.20, grid–cathode diode action limits positive excursion of K_0 to ground potential. $\therefore V_0 = 0$. From specification, $r = 0{\cdot}12$. Tentatively, put $w = 0{\cdot}05$.

(*c*) $$V_{B(nom)\,(min)} = \frac{0{\cdot}95 \times 0{\cdot}525(195 + 0) - 1{\cdot}05 \times 0{\cdot}250(195 + 40 - 100)}{0{\cdot}88 \times 0{\cdot}95 \times 0{\cdot}525 - 1{\cdot}12 \times 1{\cdot}05 \times 0{\cdot}250} = \mathbf{429V}$$

(*d*) This value of $V_{B(nom)\,(min)}$ is not too high, and so there is no need to use a smaller value of r or w.

(*e*) $$\frac{V_{B(min)}}{(1-r)} = \frac{375}{0{\cdot}88} = \mathbf{426\ V}$$

Hence $V_{B(nom)\,(min)}$ does not have to be increased above the previously determined value of 429 V.

(*f*) $$R_A = \frac{1{\cdot}12 \times 429 - 195 - 40 + 100}{0{\cdot}95 \times 0{\cdot}525} = \mathbf{691\ k\Omega}$$

(*g*) Put $R'_A = \mathbf{820\ k\Omega}$

(*h*) $$V'_{B(nom)\,(min)} = \frac{1}{0{\cdot}88}\{1{\cdot}05 \times 820 \times 0{\cdot}250 + 195 - 0\} = \mathbf{467\ V}$$

$$V'_{B(nom)\,(max)} = \frac{1}{1{\cdot}12}\{0{\cdot}95 \times 820 \times 0{\cdot}525 + 195 + 40 - 100\} = \mathbf{485\ V}$$

(*j*) $V_B = \frac{1}{2}(467 + 485) = 476$ V, say **475 V**.

(*k*) For output cathode, K_0,

$$R_{K0(min)} = \frac{12 + 0}{[0{\cdot}88 \times 475 - 195 - 0]} \cdot \frac{1{\cdot}05}{0{\cdot}95} \cdot \frac{820}{0{\cdot}85} = \mathbf{60\ k\Omega}$$

Hence put $R_{K0} = 2 \times 60 = \mathbf{120\ k\Omega}$ (no close tolerance).

(*l*) For cathodes K_1 to K_9,

$$R_{K(max)} = \frac{12 + 40 - 10}{(1{\cdot}12 \times 475 - 195 - 40 + 10)} \cdot \frac{0{\cdot}95}{1{\cdot}05} \cdot \frac{820}{0{\cdot}85} = \mathbf{123\ k\Omega}$$

$\therefore$ put $R_{K(1-9)} = \mathbf{120\ k\Omega} \pm 5\%$

Solution:

$V_B = 475\ \text{V} \pm 12\%$
$R_A = 820\ \text{k}\Omega \pm 5\%$
$R_{K(1-9)} = 120\ \text{k}\Omega \pm 5\%$ each
$R_{K0} = 120\ \text{k}\Omega \pm 20\%$

Auxiliary-anode Tubes

Two stepping tubes described by Reaney [17] are designed for direct operation of numerical indicator tubes without intermediate valves, trigger tubes, or transistors. These tubes, the GCA10G and GSA10G, are provided with ten auxiliary anodes, each situated between one of the index cathodes and the central main anode. When an index cathode is carrying current the associated auxiliary anode will carry 0–2 mA as its potential is raised from +200 to +227 V. An adjacent auxiliary anode is not so heavily primed, however, and so requires +256 to +265 V to carry the same range of current. By returning each auxiliary anode through a separate load resistor to a common supply, therefore, it is possible to obtain from each anode a negative-going pulse of the order of 40 V each time the glow rests on the associated cathode. This 40 V may be used to apply a pre-bias to one cathode of a numerical indicator tube (p. 214). By using all ten auxiliary anodes in this way the numerical indicator can be made to display the state of count held by the stepping tube. The cathode current of the numerical indicator flows through the auxiliary anode and thence to the index cathode of the stepping tube. By careful choice of gas pressure and guide bias, these tubes have been designed to operate up to 10 kc/s with cathode currents of up to 3 mA and yet maintain a life of several thousand hours even on a static count. The tube is designed to operate at constant cathode current, the current being shared between auxiliary anode and main anode as required.

Separately connected routing guides (p. 195) are provided between K_9 and K_0 so that reversible decade scalers may be constructed.

Stepping-tube Applications

Stepping tubes were developed specifically for pulse counting. The majority of applications are in this field, and may be classified into:

(*a*) Counting, read-out, and reset applications in which input pulses are counted, the state of count observed visually or electrically and a reset pulse applied after the state of count has been read [18].

(*b*) Counting to a pre-selected indication, at which an output is initiated [19]. Resetting is often performed at the same time.

(*c*) A combination of (*b*) controlling (*a*).

The first of these categories covers 'open-loop' applications: monitoring of the number of items or feet of material leaving a machine, registering the impulses from a telephone dial [12], statistical analysis of pulses into a size-distribution [20].

In the second category are batching counters [21, 22, 23] used for controlling a machine to feed pills, screws, etc., into packages in batches of any pre-selected number. Less obviously, it includes time control devices such as welder timers [24, 25] which count input pulses (often derived from the mains supply) and initiate and terminate processes at pre-determined intervals.

The third category is most familiar as the tachometer or frequency meter [26, 27]. It includes any system measuring the ratio of two mean pulse rates. One pulse input (e.g. a standard frequency oscillator) drives a batching counter, and this gates a second counter to which the second pulse input is applied.

Of these three categories, the first has already been discussed at length. The batching counter differs only in that, as well as the counting function, it is necessary to provide for an output at one or more particular states of count, and these may have to be readily adjustable. Craxton [21] has met these requirements by counting with selector tubes, outputs from chosen 'hundreds', 'tens', and 'units', cathodes being combined through a diode 'AND' gate. The gate delivers its 'AND' output as the counter arrives at the chosen number. Craxton points out that when any required number is '0' it need not be connected to the 'AND' gate. A batch of 500 thus calls only for an output from cathode K_5 of the 'hundreds' tube.

Frequency Division

The batching counter is a particular form of frequency dividing circuit. For small numbers, division by n may be effected by taking an output from a load shared by every nth index cathode of a selector tube (Fig. 7.21 (*a*)). For equally spaced output pulses the tube must have a number of cathodes divisible by n. For this purpose, a 12-way selector is useful, since division by 2, 3, 4, 6, or 12 is possible. For awkward numbers, such as 7, Hough and Ridler [2] use a transformer coupling (Fig. 7.21 (*b*)) to reset the count automatically to K_1 as soon as the glow alights on $K_{(n+1)}$.

For counts in excess of 10 or 12 various authors [2, 30, 31] have described the use of several selector tubes in cascade so that, in effect, a selector is obtained having a larger number of cathodes. Their techniques are used for pulse distribution and are outlined on pp. 205 *et seq.*

Waveform and Function Generators

Pulse train generators [28] and time markers for oscilloscopes [22, 29] are other applications of the selector tube using frequency division techniques. A staircase waveform may be generated [7] by connecting successive selector cathodes to progressively higher tappings on a

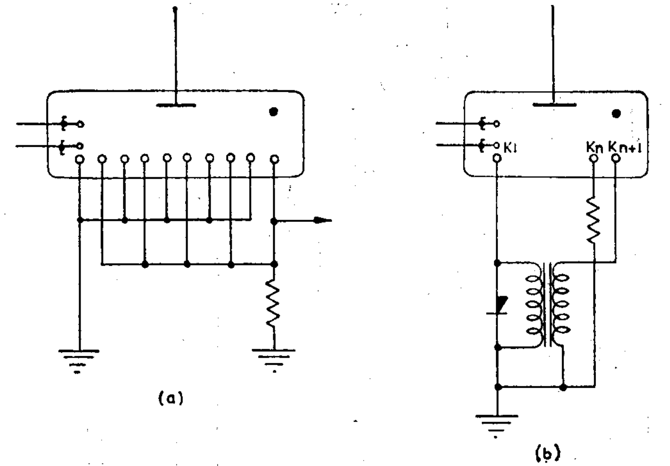


Fig. 7.21. Frequency division, (*a*) by use of common load for every nth cathode, and (*b*) by automatic resetting from $K(n+1)$ to K_1.

common cathode load. A maximum output of about 35 V is obtained from the positive end of the cathode load.

Duffy and Gilbert [30] extend this principle and, by returning each cathode to an appropriate tap on a common load, generate complex waveforms. Using several selectors in cascade, they produce waveforms composed of many small steps (Fig. 7.22). A smooth waveform is produced by generating in this way the differential of the required function and then integrating it. Resolution is improved by increasing the rate of stepping at the more complex parts of the waveform. This function generator provides outputs repeatable to about 1%.

Pulse Distributors

Selector tubes are frequently used as pulse distributors. In a telephone exchange register described by Warman and Bibb [12] the output from a selector cathode gates one of a number of circuits to which the pulse input is applied simultaneously. A variant of this [31] comprises using

diode gates to combine pulses from chosen cathodes of cascaded selector tubes. In this way two 12-way selectors can distribute pulses to 144 channels.

For smaller numbers of channels Stearman [31] has devised two methods of avoiding the limitations imposed by diode gates. Both use two or more selectors to simulate a selector having more than the usual

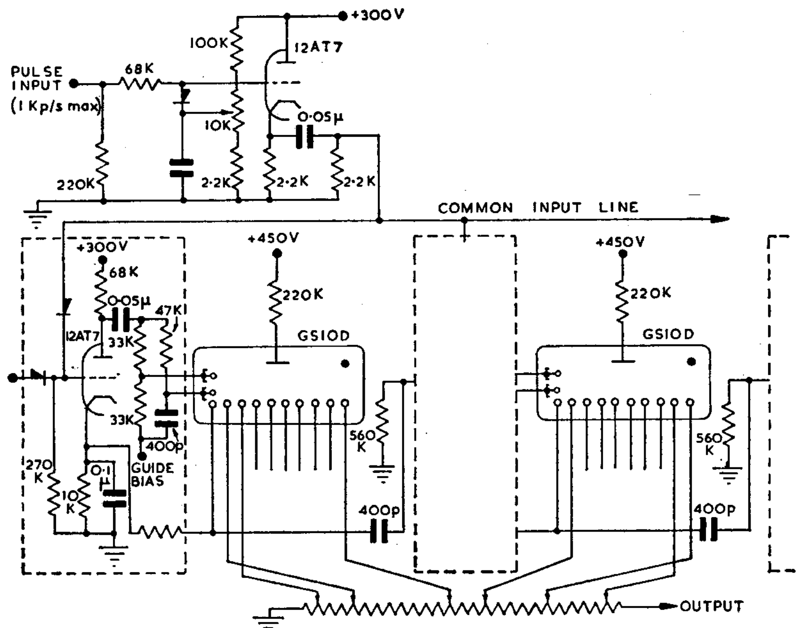


Fig. 7.22. Function generator due to Duffy and Gilbert. As the glow rests on each index cathode in turn, the output corresponds to the proportion of common load selected by the adjustable cathode taps.

10 or 12 index cathodes. In the 'glow-extinguishing method' the separate selectors share a common anode load so that only one will conduct at a time (pp. 79 and 135). Guides of the several selectors are pulsed simultaneously so that the glow advances through the cathodes K_0 to K_9 in whichever tube is conducting. As the glow leaves cathode K_9, however, a univibrator delivers a large negative pulse to cathode K_0 of the succeeding tube. This initiates a discharge to K_0 which extinguishes the glow in the tube formerly conducting. A similar coupling is provided from cathode K_9 of each tube to K_0 of the next, the tubes being connected in a ring. Stearman discusses the features of this circuit and

notes the limitation that, to produce reliable transfer from tube to tube, the univibrator pulse must have an amplitude of at least 200 V and a duration of at least 8 msec. The tubes must also be primed by ambient illumination.

For these reasons, Stearman appears to favour the alternative 'quiescent cathode method'. In this method also guide pulses are applied

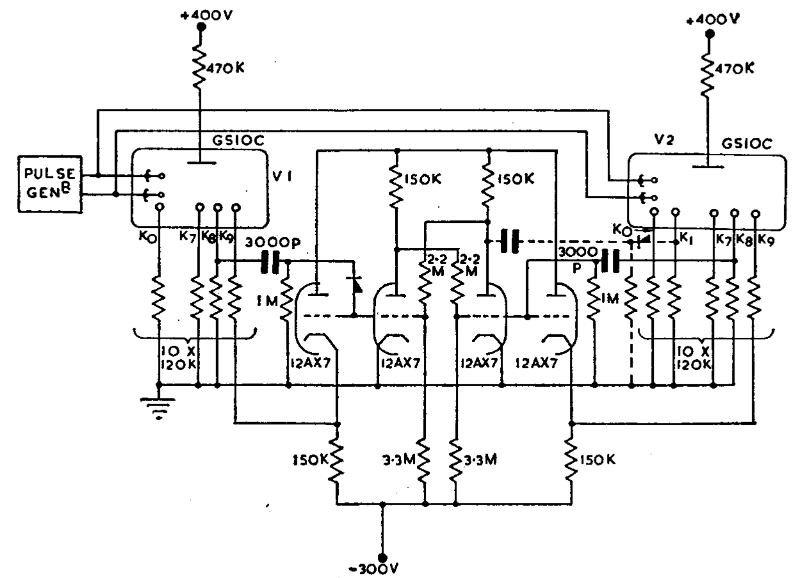


Fig. 7.23. Stearman's pulse distributor using 'quiescent cathode method'.

to all selectors in parallel. Each selector tube has a separate anode load, so all tubes conduct simultaneously. Bistable circuits apply a strongly negative bias to cathodes K_9 of all tubes but one. In these tubes the glow is thereby held on K_9 because the guide potentials never fall below the bias on K_9. In the unbiased tube, however, the glow steps in response to the guide pulses until it leaves K_8. The fall in potential of K_8 is used to reverse the state of a bistable circuit so that a pre-bias is applied to K_9 of this tube, but removed from K_9 of the succeeding tube. The guide pulses thus transfer the glow from K_9 to K_0 in the second tube. Stearman's circuit for two tubes requires only one bistable circuit and is shown in Fig. 7.23. For more than two tubes he outlines a method of arranging the bistable circuits as a ring counter. Because all tubes are continuously primed, the' quiescent cathode method' can operate more rapidly than the 'glow-extinguishing method'. Outputs cannot

be taken from the K_9 cathodes, however, and so two decade tubes give only 18 outputs. It would, however, seem practicable to obtain 20 outputs by using one 10-way and one 12-way selector. Where fewer than 18 outputs are required Stearman applies a 'reset' pulse to a pre-determined cathode (K_1 in Fig. 7.23) at the instant at which the negative bias is removed from K_9. The glow then skips to the pre-determined cathode and is almost immediately transferred by the guide pulse to the following cathode (K_2). In Fig. 7.23 the additional components indicated by broken lines cause the glow to skip cathodes K_0 and K_1 of the second selector tube so that consecutive outputs are produced at cathodes K_0 to K_8 of V_1 and then K_2 to K_8 of V_2 before returning to K_0 of V_1.

An alternative to the quiescent cathode technique is found in circuits described by Hough and Ridler [2], Duffy and Gilbert [30], and by Chao [32] wherein it is the guide pulse generators of the individual tubes which are gated to provide the operation of each tube in turn.

Read-out Systems

In data processing systems it is frequently necessary to read out the state of a decade scaler either to perform mathematical operations on the data or to transfer it to another form of storage, e.g. punched tape.

A method described by Barnes and others [33] comprises temporarily rendering the 'carry' circuits inoperative and then applying 10 guide pulse pairs to all tubes so that the glow in each performs one complete circulation. The number of guide pulses remaining after an output has been obtained from K_0 then corresponds to the indication on the tube. Similarly, the number of guide pulses occurring *before* the output pulse corresponds to the complement on 10 of the number indicated. Townsend and Camm [34] have described a calculator using a similar technique.

In some applications it may be required to use a non-destructive read-out so that, for example, a tape punch may record the state of a scaler without interrupting its operation. Chao [32] has described an appropriate technique comprising applying a carrier signal to the anode of the selector tube to be interrogated and detecting the signal at the cathode on which the glow is resting. By combining the detector outputs from like cathodes, a number of tubes may be interrogated sequentially simply by applying the carrier signal to each anode in turn.

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CHAPTER EIGHT

Register and Display Tubes

Multi-cathode display tubes may be divided into three classes:

(1) *'On-off' Indicators.* These have been developed to indicate the state of transistor circuits providing a low-voltage output.
(2) *Glow-position Register Tubes.* These give a display similar to that of a stepping tube, i.e. the glow may assume one of 10 positions, the significance of which is indicated on a surrounding escutcheon.
(3) *'Clock-face' Tubes.* This is an intermediate type, designed specifically for transistor operation. It gives a digital display, but with the numbers arranged around a 'clock face'.
(4) *Digital (and Character) Display Tubes.* These are designed for in-line digital display, all the numbers (or other characters) in a tube appearing in substantially the same position.

'On-off' Indicators

Fukukawa and Nakajo [1] have described an on–off indicator designed for use with transistor circuits. The tube has a single anode of nickel and two molybdenum cathodes so arranged that only the 'indicating' cathode is visible. A series resistor normally limits the current to about 300 μA. In the absence of pre-bias, 20 kΩ in series with the indicating cathode develops sufficient bias to ensure that most of the current flows directly to ground through the hidden 'holding' cathode. If the indicating cathode is negatively biased by 5 V or more it receives most of the anode current and glows to give an 'on' indication. Removal of the bias returns the tube to the 'off' condition.

These simple tubes may be used to indicate the state of transistor bistable circuits such as are used in computers. Although they require a supply of the order of 200 V, they dissipate only about 75 mW per tube, including the power lost in the series resistor.

Glow-position Register Tubes

Glow-position tubes resemble selector-type stepping tubes (p. 172), but have no guide electrodes. These tubes may be used to indicate the state of count of a decade scaler using thermionic valves. They may

also be used with some transistor circuits, in which application they provide a long-life display consuming little power. The high voltage for the anode supply may be derived from a transistor inverter. One use is as a display for the first stage of a decade scaler in which later stages are provided by cold cathode stepping tubes. The high-speed decade may conveniently be a trochotron tube [2].

The GR10A is representative of the glow-position register tube. It operates in the current range 50–250 μA, has a breakdown voltage, V_{IG}, of 129 V and a minimum maintaining voltage, $V_{M(min)}$, of 105 V. The tube is normally operated from a 360-V d.c. supply with 680 kΩ in series with the common anode. Thus, if all cathodes are at earth potential breakdown may occur initially to any cathode. Thereafter the anode potential falls to the maintaining voltage, V_M, (≈108 V) and no other anode–cathode gap will break down. If, however, a negative bias is applied to a non-conducting cathode breakdown will occur to this cathode when the bias is sufficiently large. The minimum required *pre-bias*, V_{PB}, is given by

$$V_{PB(min)} = V_{IG} - V_{M(min)} \tag{8.1}$$

For the GR10A, it follows that $V_{PB(min)} = 24$ V.

In use the transistor or hard-valve counter applies a negative pre-bias to each cathode in turn. In a fast decade scaler the ionization and deionization times of the register tube may be too great for the glow to follow the pre-bias. Provided the pre-bias exceeds the value given by Equation (8.1), however, the correct indication will be given almost immediately the counter stops. Thus the maximum speed of the stage depends on the counting circuit and not on the register itself.

'Clock-face' Tubes

Botden [3] has described a register tube developed for use with transistor circuits. It uses a circular wire anode of nickel spaced 5 mm from a flat annular molybdenum cathode. Ten insulated areas on the cathode ensure that any anode–cathode discharge is confined to one of the uninsulated areas. Close to each cathode and partly between it and the common anode there is a wire trigger electrode. The use of pure metal electrodes and sputtering of the cathodes in manufacture leads to a tube with closely similar characteristics in each of the anode–cathode gaps. Quite a small positive bias on one trigger will therefore ensure preferential striking if the cathode voltage is made gradually negative.

A clock-face tube is used in the circuit arrangement shown in Fig. 8.1. Half-wave rectified a.c. is applied to drive the cathode negative with

respect to ground, to which the anode and all triggers are returned. As the diode D conducts at the beginning of a new supply cycle, the trigger–cathode voltage increases towards the critical voltage, V_T. If an external circuit, X_n, applies a positive bias of a few volts to trigger T_n, the gap between T_n and cathode will break down before any other. Provided R_T is low enough to permit the flow of a trigger current of a few tens of microamperes, the discharge transfers to the anode–cathode gap.

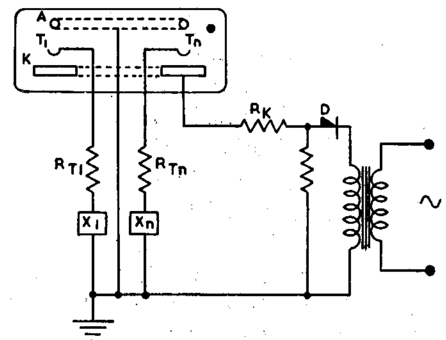

Fig. 8.1. 'Clock-face' numerical indicator tube circuit.

The cathode thereafter assumes a potential $-V_M$, where V_M is the anode–cathode maintaining voltage. As this is less than the breakdown voltage of any other trigger–cathode or anode–cathode gap, the cathode glows only opposite the positively biased trigger.

In a production form of tube, the Mullard Z550M, the anode takes the form of a mask plate in which the numbers 0–9 are perforated. The cathode glow is visible through the stencil-like perforation of the anode so that the appropriate number appears luminous and is clearly legible either in daylight or in darkness.

A bias of as little as 5 V is sufficient to produce reliable operation of a typical tube. The discharge extinguishes and re-triggers on each successive half-cycle of the supply. If the positive bias is transferred to another trigger, the discharge will move to the appropriate cathode area on the next cycle. With sinusoidal supplies the glow cannot be displaced in this way if the supply frequency exceeds 3 kc/s. This is because the 'off' half-cycle is then too short to allow the tube to deionize sufficiently for the triggers to control anode–cathode breakdown. Botden indicates that 5 V is sufficient to control the tube at frequencies of 500 c/s or less.

As in most cold cathode tubes, there is statistical delay associated with trigger–cathode breakdown (p. 62). If the trigger–cathode voltage is

rising too rapidly a delay of a few microseconds in the biased gap may allow another gap to reach triggering potential. In consequence, it is important that the rise of trigger–cathode voltage shall not exceed about 10 V/μsec, if the tube is to operate reliably without requiring excessive trigger bias. This limitation is likely to be encountered only if rectangular waveforms are applied to the cathode.

Digital (and Character) Display Tubes

Errors arise in reading meter displays, due to parallax, fatigue, and confusion of scales. These errors are not completely eliminated by displays using cold cathode stepping tubes. They are significantly reduced when an in-line digital display is used, however. An appreciation of this fact has led to a demand for clearly legible numerical displays. Of the many devices now available, the cold cathode display tube [4, 5] has proved one of the most successful. It provides excellent legibility over wide viewing angles, long life, ruggedness, economy and compatibility with thermionic valves, cold cathode tubes, or semiconductors. The display can be read even in direct sunlight, and yet the power consumption may be <500 mW per digit.

Construction

A typical tube comprises 10 nickel cathodes surrounded by a wire-mesh anode. Each cathode is in the shape of a number, letter, or symbol to be displayed. The 10 cathodes are insulated from each other and mounted about 1 mm apart, one behind the other. Separate connexions are provided to the common anode and to the individual cathodes via a multi-pin valve base or flying leads. The tube contains neon and, in long-life tubes, a trace of mercury. Its manner of operation resembles that of the register tube described above, but the cathode is identified by shape rather than position, the cathode glow enveloping the whole of the conducting cathode.

Both side-view and end-view constructions are available in a variety of sizes. End-view number tubes of about 30 mm diameter are particularly popular. Cathodes in these tubes are punched out of nickel sheet to leave a line width of about 0·4 mm. It might be thought that stacking the 10 cathodes would lead to considerable obscuration of the rearmost numbers by those in front. In practice this is not so. The conducting cathode is visible by virtue of the negative glow (p. 16), which surrounds it to a depth of about 1 mm. The total width of the glow is thus about 2 mm, and the loss of light due to obscuration by another cathode 0·4 mm wide does not exceed about 20%.

In a long-life tube the inclusion of mercury in the filling gives rise to a faint blue haze surrounding the glowing cathode. Normally the envelope is coloured orange or red to render this invisible.

Operating Conditions

Fig. 8.2 shows the simplest method of operating a digital indicator. A supply, V_B, in excess of the breakdown voltage, V_{IG}, is applied *via* a current limiting resistor, R_A, and the switch S. As shown, only the

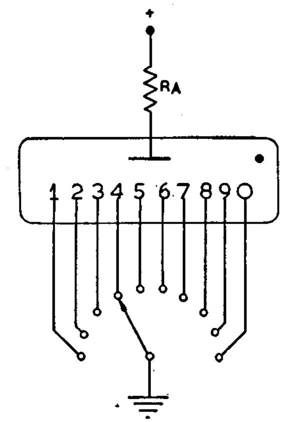

Fig. 8.2. Digital display tube operated by direct switching.

cathode '4' will carry current, the other cathodes assuming a potential intermediate between those of the anode and the active cathode. R_A must be chosen with some care. Sufficient current must be passed to fill the whole of any cathode. If this is exceeded, however, the tube operates in the region of abnormal glow (p. 15). With further increase of current, cathode sputtering increases rapidly and severely reduces tube life. For good operation throughout life it is therefore important that the manufacturer's recommendations be strictly observed. A new tube will generally perform satisfactorily on rather less than the specified minimum current. After some time, however, material sputtered from adjacent cathodes locally contaminates some of the cathodes so that the work function is increased over the contaminated areas. The current flowing to a cathode then tends to be carried only by the clean areas. It is not until these areas are running severely into the abnormal glow region that the maintaining voltage increases sufficiently to cause the glow to extend over the contaminated parts of the cathode. Accordingly,

the manufacturer specifies a minimum current which will cause the glow to fill a cathode even when it is somewhat contaminated.

Life

From the foregoing it will be appreciated that tube life is usually determined by sputtering of cathode material. This can produce gradual tube failure by one or more of the following effects:

(1) Blackening of the glass envelope.
(2) Some numbers may not be fully illuminated, due to local cathode contamination.
(3) Parts of a cathode may be eroded completely.
(4) Sputtered material may bridge insulators so that several cathodes tend to light at once.

Tube life is shortest if the discharge rests continuously on one cathode. If all the cathodes are used in turn and at sufficiently frequent

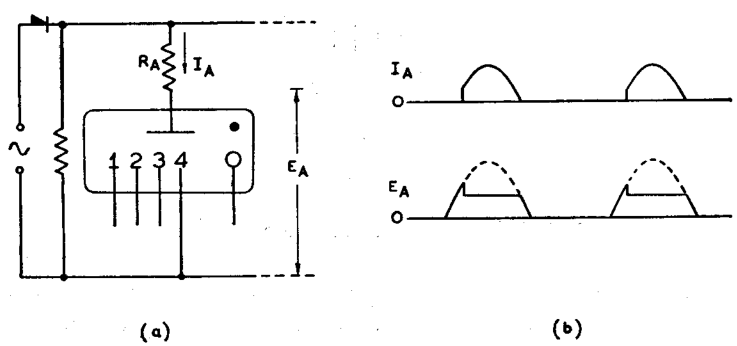

Fig. 8.3. (*a*) Numerical display tube operated on unsmoothed, rectified a.c. (*b*) Corresponding anode current and anode voltage waveforms.

intervals material sputtered on to a cathode is to a large extent removed again by further sputtering so that much of it is deposited elsewhere.

In early designs of numerical indicator tubes a life of only about 1,000 hours was obtained from a single cathode operating continuously on a d.c. supply. With the display regularly cycled through each cathode in turn, the life increased to about 5,000 hours. 'Long-life' tubes are now available in which a trace of mercury is included in the filling. This gives a greatly increased life: some 5,000 hours on a single cathode, rising to 30,000 hours on a display using each cathode in turn.

Still longer lives are possible when pulsed h.t. supplies are employed. A commonly used arrangement (Fig. 8.3) comprises operating the tube

on unsmoothed half-wave rectified a.c. In this way the tube life is trebled, since it is conducting only for about one-third of the total cycle. An extension of this principle is shown in Fig. 8.4, wherein a trigger tube extinguishes the numerical indicator before the supply voltage reaches its peak value. The mode of operation is comparable with that

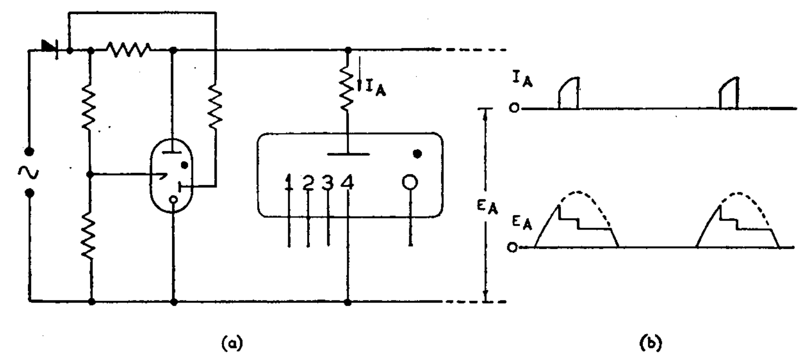

Fig. 8.4. (*a*) Trigger tube used to reduce duty factor and extend life of numerical indicator tube. (*b*) Corresponding anode current and voltage waveforms for display tube.

of the stabilizing circuit of Fig. 5.4. Several tubes may be controlled by a single trigger tube, the life of which may be arranged to match the life of the indicator tubes [6]. In using these techniques to extend tube life, it must be remembered that the luminance of the display falls in the same ratio as the mean current. If the tube current is reduced to 10% to produce a ten-fold increase in tube life, its luminance will be reduced to one-tenth. The reduction in luminance is not likely to be serious unless the display is to be read in bright sunlight.

Operation

Over the working current range the anode–cathode voltage, V_M, of a number tube is almost constant. Suitable values and tolerances for supply voltage, V_B, and anode resistor, R_A, may therefore be arrived at using the design procedure for the stabilizer diode circuit (p. 26 *et seq*). Design is here simplified by the absence of a shunt load current, I_L. (Hence $I_{L(min)} = I_{L(max)} = 0$.)

For some tubes, it is recommended equalizing resistors be connected in series with certain cathodes (e.g. '1' and '7') having areas significantly less than others and accordingly requiring smaller currents. Suitable values may then be obtained as a difference between the calculated

values of R_A required for the larger and smaller cathode currents. Most modern tubes do not require equalizing resistors.

For tubes operating on half-wave rectified a.c., Fig. 5.7 is applicable if k is interpreted as $V_m \sqrt{2}/V_M$. Authorities differ in their approach to the design of circuits for half-wave operation. McLoughlin [4] explicitly recommends an anode resistor chosen to restrict the peak cathode current to the maximum value permitted with d.c. operation. The mean current will then be about a quarter of this value, and tube life may be

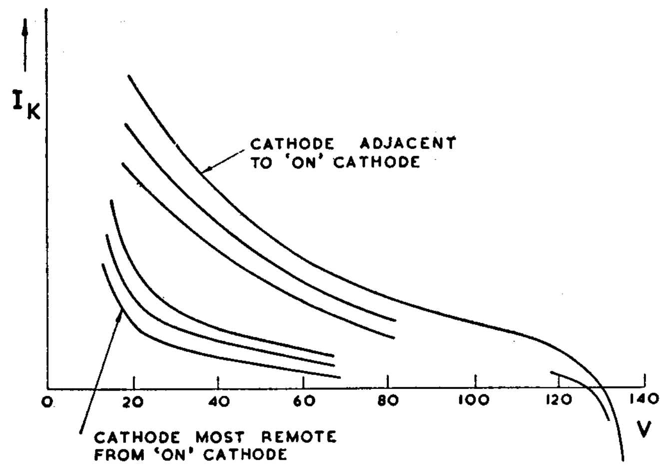

Fig. 8.5. Current-sharing between 'off' cathodes in numerical indicator tube as function of pre-bias voltage between 'off' and 'on' cathodes.

expected to be increased considerably. McDougall [7] advises a relatively low value of series resistor, and thereby obtains a mean current comparable with that recommended for smoothed-d.c. operation. This will cause the glow to cover the cathode fully over the greater part of the period of conduction. Sputtering is increased by the increased peak current, but, on the other hand, full use of the cathode area helps to preserve cathode cleanliness. Consequently, tube life may be no worse in this mode of operation, and cathode luminance is certainly higher. McDougall claims that with low-pre-bias voltages the display contrast is actually improved when an unsmoothed rectified a.c. supply is used.

The display contrast is a function of current-sharing between cathodes. Fig. 8.5 shows that current-sharing is reduced as pre-bias is increased. Typically, display contrast is acceptable at 40 V bias and improves up to within 20 V of the anode maintaining voltage, i.e. up to about 120 V.

The close spacing of the cathodes leads to considerable priming of

adjacent anode–cathode gaps by the discharge. Consequently, Equation (8.1) is not applicable to the numerical indicator tube.

If the tube is operating in light, initial priming of an anode–cathode gap is effected by photo-emission. In complete darkness a delay of the order of one second may arise due to statistical variations in natural radiation on which priming is then dependent. This may be troublesome if the display is to be switched on only briefly, e.g. for photographic recording. One solution is to illuminate the tube by a small tungsten lamp covered by a violet filter (Ilford 621) and to render this priming illumination invisible to camera and/or eye by viewing through a yellow filter (Ilford 110).

With certain tubes, e.g. the Ericsson GR10G, a simpler solution may be used. A continuous priming discharge of a few microamperes is established between a pair of adjacent pins which pass through the glass base of the tube, but are not connected to any of the electrodes. This discharge produces a faint glow below the lower ceramic and is therefore readily masked off. The pins used should preferably be remote from any connected to internal electrodes. A limiting resistance of 10–100 MΩ must be inserted in series with one lead connecting the chosen pins to a supply $\geqslant$225 V d.c. or 200 V r.m.s. a.c.

Tubes operating on half-wave rectified power supplies present no special problems when used in complete darkness. Once a discharge has been established, the deionization time is long enough to ensure reliable restriking on each successive positive half-cycle.

Design Procedure for Character Display Tube Operating on Rectified (Unsmoothed) A.C. Supply (*Fig.* 8.3) (also applicable to symmetrical diode indicators on a.c. supplies)

(*a*) Set out the following data for the selected tube:
- Anode breakdown potential, V_{IG}
- Anode maintaining potential, V_M
- Maximum permissible average cathode current, $I_{K(av)\,(max)}$
- Minimum permissible average cathode current, $I_{K(av)\,(min)}$
- Maximum permissible peak cathode current, $I_{K(pk)\,(max)}$

(*b*) Set out further design data:
- R.m.s. voltage of a.c. supply, V_m
- Fractional tolerances, $+r_1$ and $-r_2$, on a.c. supply voltage
- Full-wave rectification ($u = 1$), or half-wave ($u = \frac{1}{2}$)
- (For indicators operating on unrectified a.c., $u = 1$.)

(*c*) Check that the a.c. supply voltage is sufficiently high to ensure striking, i.e.,

$$V_m \geqslant \frac{V_{IG}}{(1 + r_1)\sqrt{2}} \tag{5.88a}$$

(*d*) Determine *k*.

$$k = \frac{V_m\sqrt{2}}{V_M}$$

(*e*) From Fig. 5.7, determine the corresponding value of *p*.

(*f*) Determine $R_{A(min)}$

$$\frac{(1 + r_1)V_m\sqrt{2} - V_M}{I_{K(pk)\ (max)}} \leqslant R_{A(min)} \geqslant \frac{u}{p} \cdot \frac{(1 + r_1)V_m\sqrt{2} - V_M}{I_{K(av)\ (max)}}$$

(*g*) Determine the maximum factor, *g*, by which $I_{K(av)\ (min)}$ must be increased for design purposes.

V_{IG}/V_m	g
0·75	1·2
0·95	1·4
1·10	1·7
1·20	2·0

(*h*) Determine $R_{A(max)}$.

$$R_{A(max)} = \frac{u}{p} \cdot \frac{(1 - r_2)V_m\sqrt{2} - V_M}{g \,.\, I_{K(av)\ (min)}}$$

(*j*) Determine R_A.

$$R_A \approx \tfrac{1}{2}(R_{A(min)} + R_{A(max)})$$

(*k*) Choose a preferred value for R_A and a tolerance conforming with the limits given by (*f*) and (*h*) above.

EXAMPLE 8.1 Character Display Tube on Rectified (Unsmoothed) A.C. Supply

Determine the value and tolerance of the anode load resistance, R_A, for a Z520M numerical indicator tube operating on half-wave rectified a.c. supply of 240 V r.m.s., +10%, −15%.

(*a*) $V_{IG} = 170$ V $\qquad I_{K(av)\ (max)} = 2{\cdot}5$ mA

$V_M = 140$ V $\qquad I_{K(av)\ (min)} = 1{\cdot}0$ mA

$I_{K(pk)\ (max)} = 10$ mA

(*b*) $V_m = 240$ V r.m.s. $\qquad r_1 = 0{\cdot}10$

$u = \tfrac{1}{2}$ $\qquad r_2 = 0{\cdot}15$

(*c*)
$$\frac{V_{IG}}{(1 + r_1)\sqrt{2}} = \frac{170}{1{\cdot}10\sqrt{2}} = \mathbf{109\ V}\ (< V_m)$$

(*d*)
$$k = \frac{240\sqrt{2}}{140} = 2{\cdot}42$$

(*e*) From Fig. 5.7, $p = 1{\cdot}94$.

(*f*)
$$\frac{(1 + r_1)V_m\sqrt{2} - V_M}{I_{K(pk)\ (max)}} = \frac{1{\cdot}10 \times 240\sqrt{2} - 140}{10} = \mathbf{23{\cdot}4\ k\Omega}$$

$$\frac{u}{p} \cdot \frac{[(1 + r_1)V_m\sqrt{2} - V_M]}{I_{K(av)\ (max)}} = \frac{\tfrac{1}{2}}{1{\cdot}94} \times \frac{(1{\cdot}10 \times 240\sqrt{2} - 140)}{2{\cdot}5} = \mathbf{24{\cdot}1\ k\Omega}$$

$$\therefore\ R_{A(min)} = \mathbf{24{\cdot}1\ k\Omega}$$

(*g*)
$$V_{IG}/V_m = 170/240 = 0{\cdot}71$$

$$\therefore\ g = 1{\cdot}2$$

(*h*)
$$R_{A(max)} = \frac{\tfrac{1}{2}}{1{\cdot}94} \times \frac{(0{\cdot}85 \times 240\sqrt{2} - 140)}{1{\cdot}2 \times 1{\cdot}0} = \mathbf{31{\cdot}8\ k\Omega}$$

(*j*) $R_A \approx \tfrac{1}{2}(24{\cdot}1 + 31{\cdot}8) = 27{\cdot}9$ kΩ, say **27 kΩ.**

(*k*) 24·1 kΩ is 10·5% below 27 kΩ. Hence use a 10% tolerance.

Solution:

In the circuit of Fig. 8.3,

$$R_A = 27\ \text{k}\Omega\ \pm 10\%$$

Thermionic Tube Drive

Fig. 8.6 represents a straightforward, if rather cumbersome, way of producing an adequate pre-bias on a numerical indicator tube. It comprises returning each cathode to the anode of an individual thermionic triode. All triodes are biased to cut-off except the one which is to pre-bias the required number cathode. In Fig. 8.6 a selector stepping tube removes bias from the appropriate triode so that the number tube indicates the position of the glow in the stepping tube. With the component values shown, a second numerical indicator tube may, if required, be connected to the same triode anodes. Thus a remote indicator calls only for an additional tube and 27-kΩ anode resistor.

It is possible to obtain a 'carry' pulse from the anode of the triode connected to K_0. If this is done, a shunt resistor, R_0, is required so that the leading edge of the pulse shall be independent of the ionization time of the number tube.

Transistor Drive

Space and power may be saved by using *n–p–n* transistors to replace the thermoinic tubes of Fig. 8.6. A high-voltage transistor is required to

provide the required pre-bias of 40–50 V. Somlyody [8] has shown that the normal collector breakdown voltage of a transistor is increased if a reverse bias of at least 200 mV is applied to the base-emitter junction. As a result it is possible to use in Fig. 8.7 certain transistors (e.g. XA701, 2N585, 2S701) for which the maximum collector–emitter voltage rating is normally 20–25 V.

A transistor is bottomed when in the 'on' condition. The value of

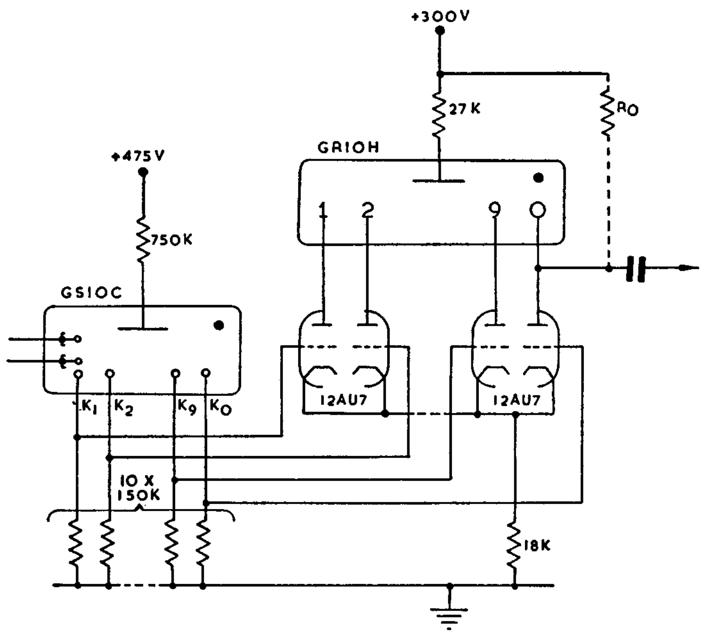

Fig. 8.6. Thermionic tube drive of numerical indicator from selector stepping tube.

R_A is thus determined exactly as for a directly switched tube. It is only for the determination of transistor base return resistor, R_B, that a special design procedure [9] is required.

If R_B is returned to a negative bias, $-V_X$, then when the transistor is in the 'off' condition,

$$R_{B(max)} = \frac{V_X - V_{be(off)}}{I_{co(max)}} \tag{8.2}$$

In the 'on' condition,

$$R_{B(min)} = \frac{V_X + V_{be(on)}}{I_{in(min)} - I_b} \tag{8.3}$$

where $V_{be(on)}$ = base-emitter voltage corresponding to base current, I_b, and a collector current, I_c, equal to $I_{A(max)}$ of the number tube.

In Fig. 8.7 R_B is the resistor in each cathode lead of the selector tube and I_m is the current in that cathode. In a representative case $R_B \approx 10$ kΩ. Somlyody [8] describes the 'Trixie' modules of ten transistors sold by Burroughs specifically to switch number tubes. In these modules the

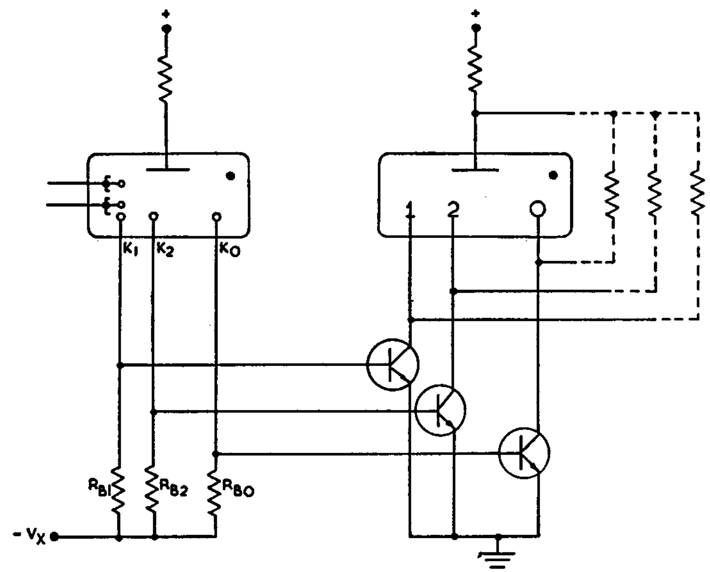

Fig. 8.7. *N-p-n* transistors used to couple numerical indicator tube to selector stepping tube.

working range of temperature is extended by connecting 1·5 MΩ in parallel with each anode–cathode gap of the number tube.

An Ericsson publication [10] shows how a digital indicator tube may be used to provide a decimal read-out from a transistor binary counter. The arrangement is shown in Fig. 8.8. A diode matrix provides binary-decimal conversion and *n-p-n* transistors switch the number-tube cathodes.

Drive by Shockley Diode Ring

P-n-p-n diodes (Shockley diodes) may also be used to switch digital display tubes. They can conveniently switch the rather large pre-bias voltages required by some tubes (e.g. the GR10G) and have also current-carrying capacity adequate for the switching of several tubes at once.

The Shockley diode [11] may be regarded as the semiconductor equivalent of the 'difference diode' (p. 46). A typical Shockley diode has a breakover voltage of 100 V $\pm$ 10 V and shows a forward voltage drop of $<$1 V when conducting. Used in the circuit of Fig. 8.9, *p–n–p–n* diodes

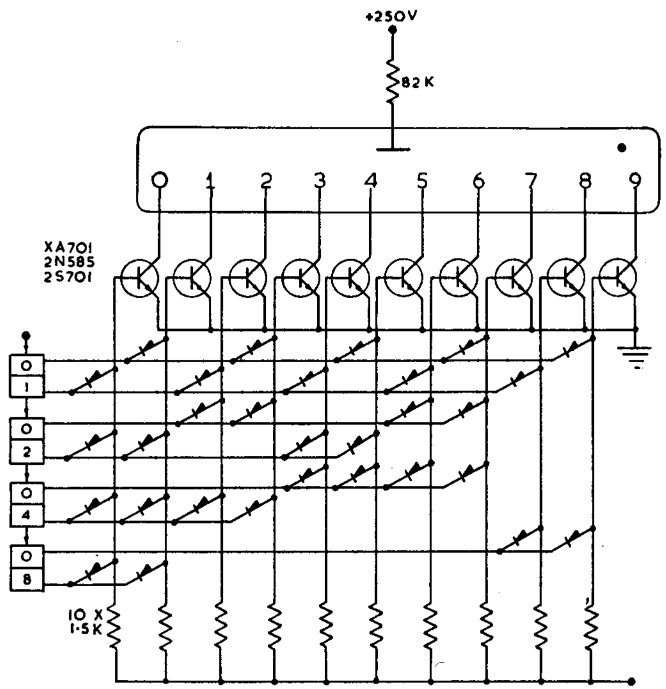

Fig. 8.8. Decimal indication from a binary scaler. The diode matrix provides binary-to-decimal conversion and the output is applied to the numerical display tube by *n–p–n* transistors.

provide not only switching of the display tube, but counting also. The diode ring counter of Fig. 8.9 operates on the same principle as that of Fig. 3.14. The display tubes may be switched on and off (by raising and lowering their anode supply voltages) without disturbing the state of the Shockley diode counter.

Trigger Tube Drive

By returning each cathode of a number tube directly to the anode of a separate trigger tube, cold cathode driving circuits may be constructed capable of performing a variety of functions.

(a) Ring Counter

Hodgson [12] has described a ring counter using the Z70U/Z700U to count at up to 2·5 kc/s and drive a Burroughs 'Nixie' HB106 number

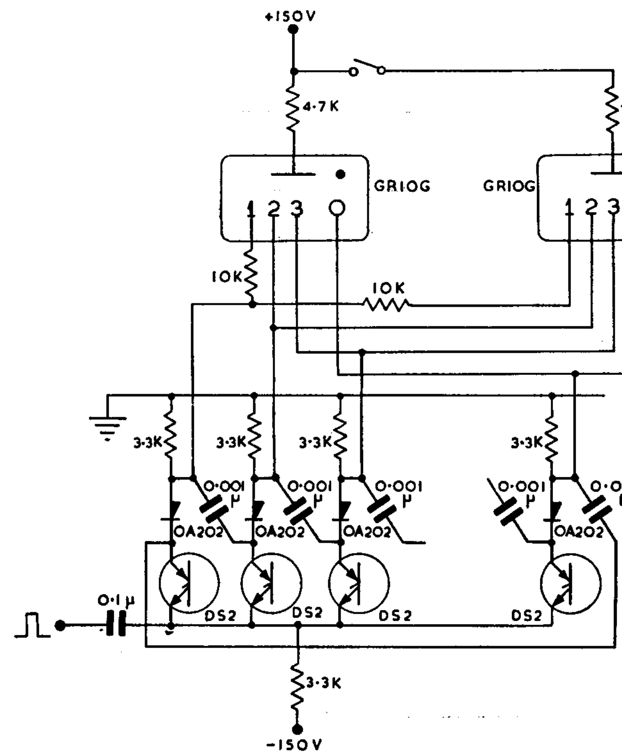

Fig. 8.9. Shockley diode ring counter driving two numerical indicator tubes, one having switched anode supply.

tube. Fig. 8.10 shows a very similar circuit published by Mullard [13]. A Z700U, on the left of the diagram, operates in the self-quenching mode to provide low impedance positive-going pulses. These are applied, through 100 pF capacitors, to the triggers of ten tubes connected to form a ring counter (p. 135). In addition to the common 150 kΩ anode resistor, a further 68 kΩ is connected in series with each of the ten anodes. The conducting tube develops a pre-bias for the Z520M across one of these 68-kΩ resistors. The 0·01 μF capacitors shunting the 68-kΩ anode resistors make the maximum counting speed of the ring less dependent on the rather long ionization and deionization times of the Z520M.

(b) Digital Display from Selector Tube

An Ericsson circuit [10] provides a simple 'add-on' digital display which may be connected to an existing stepping-tube counter using selector tubes or to a transistor circuit which provides an output of 12 V or more. Fig. 8.11 shows that GR10K and GTE120Y tubes operate in series across a half-wave rectified a.c. supply. The tubes therefore extinguish

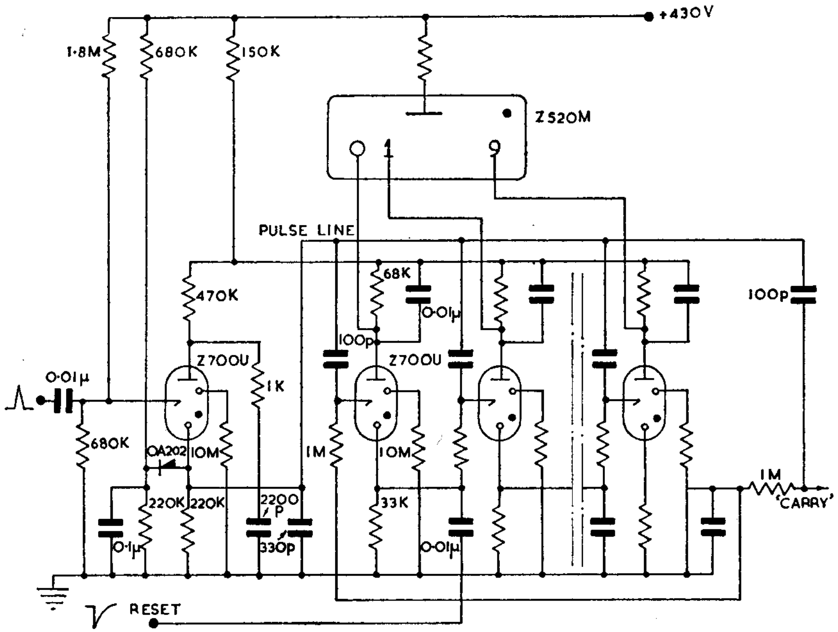

Fig. 8.10. Trigger tube ring counter driving numerical display tube directly.

on alternate half-cycles of the supply. As the rectified voltage across the tubes increases with a new cycle, no tube strikes until the GTE120Y cathodes are nearly 120 V below ground. A positive trigger bias of as little as 12 V will then ensure that a particular tube fires before the others. As soon as this tube conducts, the cathodes of all GTE120Y tubes rise by virtue of the voltage drop across the 82-kΩ series resistor. It follows that no other trigger tube can fire. With further increase in rectified voltage the drop across the 100-kΩ anode resistor of the trigger tube rises to 150 V and the GR10K strikes to the appropriate cathode.

The above process is repeated at the beginning of each new cycle of the mains supply. If the state of the input changes a new number will be displayed by the GR10K when a new cycle begins.

(c) Storage Display

Using a d.c. supply instead of rectified a.c., a storage display results. Once one GTE120Y has struck to pre-bias the GR10K in Fig. 8.12, the input has no further control over the display. When contacts *a*/1 and *a*/2 are open the tubes extinguish. When they reclose, therefore, the display is re-established, corresponding to the new input.

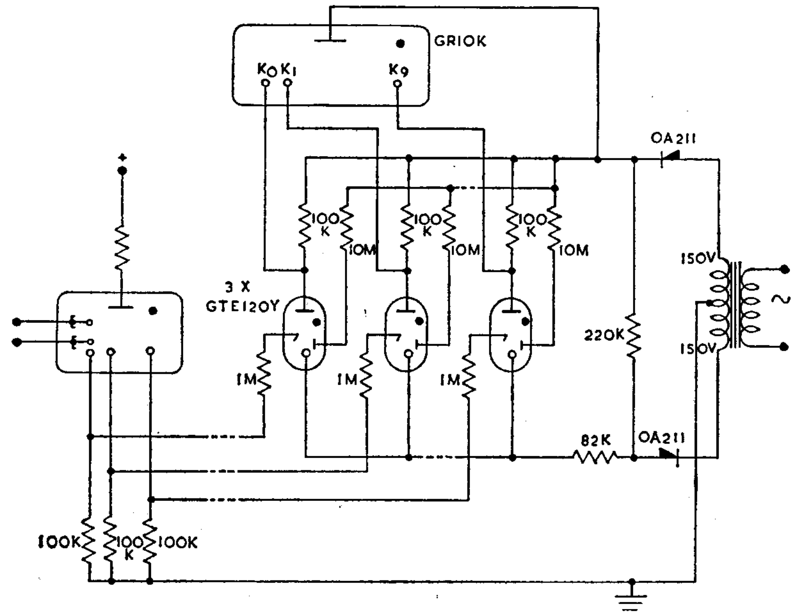

Fig. 8.11. 'Add-on' trigger tube circuit coupling numerical display tube to selector stepping tube.

Given suitable switching of contacts *a*/1 and *a*/2 in each unit, a number of units similar to Fig. 8.12 may be used to allow a single frequency meter or digital voltmeter to operate in time-division multiplex, displaying several inputs simultaneously.

Trochotron Drive

The trochotron [2] is a high-vacuum stepping tube having a thermionic cathode and capable of operating at up to 2 Mc/s. It is particularly suited to the switching of large digital display tubes (Fig. 8.13), each of the 10 target anodes being able to carry up to 18 mA. The anodes show pentode-type characteristics and, accordingly, the trochotron is

not easily affected by the relatively long ionization and deionization times of the gas-filled display tube. In fact, the trochotron can still operate at its maximum frequency provided the target potential does not fall below the knee of the characteristic. At stepping speeds above

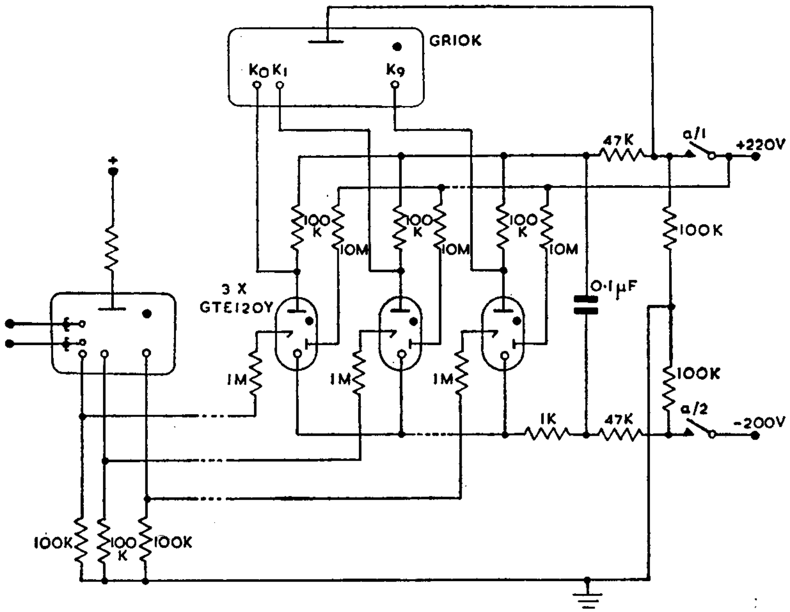

Fig. 8.12. 'Storage' display unit using trigger tubes. The numerical display tube indicates the state of count of the selector stepping tube at the instant of closure of $a/1$ and $a/2$.

300 kc/s the display tube will fail to ionize and the 12-kΩ resistors in Fig. 8.13 are accordingly needed to provide a path for the target current. As the trochotron stepping slows down or stops, the display tube will ionize once more. Even when counting slowly, however, the stepping action of the trochotron is so abrupt that the display tube may fail to conduct to the newly-switched cathode for some 200–300 µsec.

The value of target resistor, R_t, in Fig. 8.13 will give the correct display tube current if:

$$R_t = \frac{V_t - V_B + I_A R_A + V_M}{I_t - I_A} \qquad (8.4)$$

As the knee in the target characteristic occurs at approximately half

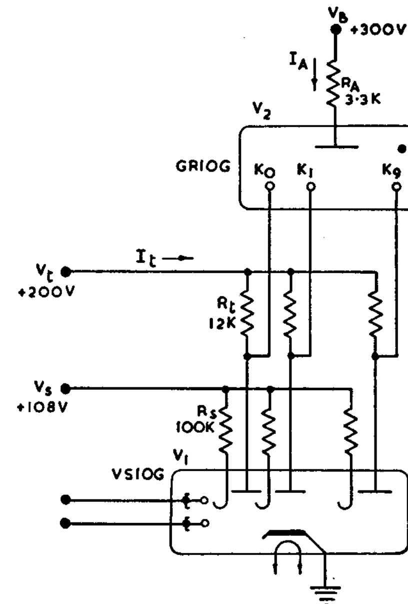

Fig. 8.13. Trochotron tube drive of numerical display tube.

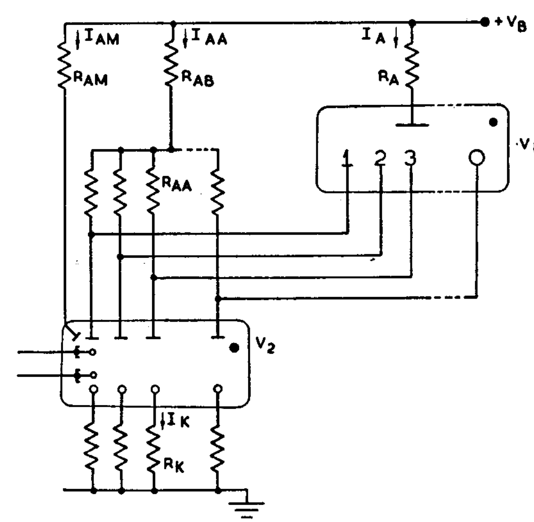

Fig. 8.14. Numerical display tube operated directly by auxiliary-anode stepping tube.

the spade supply voltage, V_s, bottoming of the trochotron will be avoided if:

$$R_t < \frac{V_t - \frac{1}{2}V_s}{I_t} \tag{8.5}$$

Although the pentode-type characteristics of the trochotron targets make an anode resistor not strictly necessary for the display tube, it is nevertheless desirable. If the drop across R_A is made of the same order as that across R_t, then the dependence of I_A upon I_t is usefully reduced.

Direct Drive by Auxiliary-anode Stepping Tube

The auxiliary-anode stepping tube (p. 203) described by Reaney [14] was developed specifically to operate a digital display tube, using the circuit of Fig. 8.14.

A decade scaler using auxiliary-anode tubes for counting can therefore be provided with digital display at very little extra cost. Close-tolerance resistors and well-regulated supplies will be required, however, if tube ratings are to be observed strictly.

Design of the anode–cathode circuit begins with the choice of cathode resistor R_K. The manufacturer recommends a maximum value such that $I_K R_K \not> 10$ V. If larger output voltages are required they are obtainable (with the opposite polarity) at the auxiliary anodes.

Writing V_M, V_{MM}, and V_{MA} for the maintaining voltages of the display tube and the main anode- and auxiliary anode–cathode gaps of stepping tube respectively, one can write the following equations:

$$V_B = I_K R_K + V_{MA} + V_M + I_A R_A \tag{8.6}$$

$$V_B = I_K R_K + V_{MM} + I_{AM} R_{AM} \tag{8.7}$$

$$V_B = I_K R_K + V_{MA} + I_{AA}(R_{AA} + R_{AB}) \tag{8.8}$$

Also, to ensure that no significant current is carried either by the ‘off’ cathodes of the display tubes or by the ‘off’ auxiliary anodes of the stepping tube, R_{AA} is chosen to drop the maximum voltage ($\gtrsim$40 V) which can be applied between a conducting auxiliary anode and an adjacent non-conducting anode. Thus,

$$I_{AA} R_{AA} \gtrsim 40 \text{ V} \tag{8.9}$$

Provided I_{AA} is small, so that R_{AA} is large, this value is not very critical: any unwanted currents have to flow through one of the R_{AA} resistors, and in so doing produce sufficient auto-bias to reduce the unwanted current considerably.

Finally,

$$I_K = I_{AM} + I_{AA} + I_A \tag{8.10}$$

Rewriting these equations to correspond to maximum and minimum values, fully toleranced values of R_{AM}, R_{AA}, R_{AB}, R_A, and V_B may be determined. The following indicates the basis of the design procedure beginning on p. 233.

From Equation (8.6),

$$I_A = \frac{1}{R_A}(V_B - V_{MA} - V_M - I_K R_K) \tag{8.6a}$$

$$\therefore \quad \frac{dI_A}{dV_B} = \frac{1}{R_A} = \frac{I_A}{V_B - V_{MA} - V_M - I_K R_K} \tag{8.11}$$

Assuming the greater part of I_K is due to I_A, write:

$$\frac{\delta I_K}{I_K} \approx \frac{\delta I_A}{I_A} = \frac{1}{I_A} \cdot \frac{dI_A}{dV_B} \cdot \delta V_B$$

Substituting from Equation (8.11)

$$\frac{\delta I_K}{I_K} \approx \frac{\delta V_B}{V_B - V_{MA} - V_M - I_K R_K} \tag{8.12}$$

where $\delta I_K = I_{K(max)} - I_{K(min)}$

and $\delta V_B = V_{B(max)} - V_{B(min)} = 2rV_B$

r = fractional tolerance on V_B

Equation (8.12) may be re-written to give

$$r \approx \frac{1}{2} \cdot \frac{\delta I_K}{I_K} \left\{ 1 - \frac{V_{MA} + V_M + I_K R_K}{V_B} \right\} \tag{8.12a}$$

It will be seen, therefore, that $r < \frac{1}{2}\delta I_K / I_K$ and, unless very high values of V_B are used, r will be considerably smaller than this limiting value.

With practical values, $(V_{MA} + V_M) \approx 0{\cdot}8 V_B$,

$$\therefore \quad r \approx \frac{1}{10} \cdot \frac{\delta I_K}{I_K} \tag{8.12b}$$

Using Equation (8.12b) to choose a reasonable value of r, the corresponding minimum supply voltage, V_B, may be determined from Equation (8.12a).

This design procedure is somewhat pessimistic in that it neglects the regulation of V_M and V_{MA} with current changes and also the fact that a significant proportion of the current passes through resistors effectively in shunt with the display tube. On the other hand, no allowance has yet been made for resistor tolerances.

Allowance should strictly be made also for tolerances on the maintaining voltages of the various anode–cathode gaps in the two tubes. In

practice, however, it is reasonable to neglect these: over the range of working current the incremental resistance of the discharge produces a change in V_M which is comparable with the manufacturer's tolerance on V_M. Thus by neglecting the favourable effect of the incremental resistance, one makes some allowance for the (unfavourable) tolerance on V_M.

With the value of supply voltage, V_B, and its fractional tolerance, r, settled, the designer may next determine the anode load, R_A, of the display tube and its tolerance w_A.

At this stage the maximum and minimum values of $(I_{AM} + I_{AA})$ are calculated, corresponding to the maximum and minimum values of I_A and I_K which will arise.

As $V_{MA} \approx V_{AA}$, the tolerance w_M on R_{AM} may be estimated by assuming $(R_{AA} + R_{AB})$ is in parallel with R_{AM}. Then,

$$(I_{AM} + I_{AA})_{(max)} \approx \frac{(1 + r)V_B - V_{MM} - I_K R_K}{(1 - w_M)R_P} \qquad (8.7a)$$

and

$$(I_{AM} + I_{AA})_{(min)} \approx \frac{(1 - r)V_B - V_{MM} - I_K R_K}{(1 + w_M)R_P} \qquad (8.7b)$$

where

$$R_P = \frac{R_{AM}(R_{AA} + R_{AB})}{R_{AM} + R_{AA} + R_{AB}}$$

Combining Equations (8.7a) and (8.7b),

$$\frac{1 + w_M}{1 - w_M} \approx \frac{(I_{AM} + I_{AA})_{(max)}}{(I_{AM} + I_{AA})_{(min)}} \cdot \left\{\frac{(1 - r)V_B - V_{MM} - I_K R_K}{(1 + r)V_B - V_{MM} - I_K R_K}\right\}$$

whence

$$w_M \approx \frac{1}{2} \cdot \left[\frac{(I_{AM} + I_{AA})_{(max)}}{(I_{AM} + I_{AA})_{(min)}} \cdot \left\{\frac{(1 - r)V_B - V_{MM} - I_K R_K}{(1 + r)V_B - V_{MM} - I_K R_K}\right\} - 1\right] \qquad (8.13)$$

The value of the stepping tube main anode load resistor, R_A, and its tolerance, w_M, may thus be found, leaving until last the fixing of R_{AA} and R_{AB} and their tolerances. Because this is the last step in the design, the tolerance on $(R_{AA} + R_{AB})$ is likely to be small. The individual values of R_{AA} and R_{AB} are much less critical, however, and it is thus possible to meet the design requirements using preferred value resistors. Moreover, as R_{AA} is only a fraction of the value of R_{AB}, a correspondingly larger tolerance may be permitted in R_{AA}.

Design Procedure for Display Tube Directly Driven by Auxiliary-anode Stepping Tube

(*a*) Set out the following data for the auxiliary-anode tube:
- Maximum cathode current, $I_{K(max)}$
- Minimum cathode current, $I_{K(min)}$
- Main anode–cathode maintaining voltage, V_{MM}
- Auxiliary anode–cathode maintaining voltage, V_{MA}
- Maximum main anode current, $I_{AM(max)}$
- Minimum main anode current, $I_{AM(min)}$
- Value of cathode resistor, R_K

(*b*) Set out the following data for the character display tube:
- Maximum anode current, $I_{A(max)}$
- Minimum anode current, $I_{A(min)}$
- Anode–cathode maintaining voltage, V_M.

(*c*) Evaluate

$$\frac{\delta I_K}{I_K} = \frac{I_{K(max)} - I_{K(min)}}{\frac{1}{2}(I_{K(max)} + I_{K(min)})} = A$$

(*d*) Tentatively setting r, the fractional tolerance on supply voltage, V_B, such that $r \approx \frac{1}{10} \cdot \frac{\delta I_K}{I_K}$, calculate $V_{B(nom)\,(min)}$,

$$V_{B(nom)\,(min)} \approx \frac{A}{A - 2r}(V_{MA} + V_M + I_K R_K) \qquad (8.12c)$$

where

$$I_K = \tfrac{1}{2}(I_{K(max)} + I_{K(min)})$$

(*e*) If this value of $V_{B(nom)\,(min)}$ is not excessive, choose a convenient value of $V_{B(nom)}$ not less than the calculated value.

If the value of $V_{B(nom)\,(min)}$ is too high, repeat step (*d*) using a smaller value of r.

(*f*) Determine the nominal anode load, R_A, for the display tube.

$$R_A \approx \frac{(V_B - V_{AA} - V_M - I_K R_K)}{\frac{1}{2}(I_{A(max)} + I_{A(min)})} \qquad (8.6b)$$

Round off to the nearest preferred value.

(*g*) From Relation (8.14), determine the fractional tolerance, $w_{A(max)}$, on R_A which leaves I_A influenced equally by w_A and by r.

$$w_{A(max)} \approx \frac{r \,.\, V_B}{V_B - V_{MA} - V_M} \qquad (8.14)$$

(*h*) Tentatively choose a value of $w_A < w_{A(max)}$

(*j*) Determine the maximum value, $I_{A(max)}$, of I_A which will occur with tolerances r and w_A

$$I_A'{}_{(max)} = \frac{(1 + r)V_B - V_{MA} - V_M - I_K R_K}{(1 - w_A)R_A} \quad (8.6c)$$

(*k*) Check $I_A'{}_{(max)} < I_{A(max)}$. If this relation is not satisfied, choose a smaller tolerance w_A and repeat steps (*j*) and (*k*).

(*l*) Determine the minimum value, $I_A'{}_{(min)}$, of I_A.

$$I_A'{}_{(min)} = \frac{(1 - r)V_B - V_{MA} - V_M - I_K R_K}{(1 + w_A)R_A} \quad (8.6d)$$

(*m*) Check $I_A'{}_{(min)} > I_{A(min)}$. If this relation is not satisfied, choose a smaller tolerance w_A and repeat steps (*j*), (*l*), and (*m*).

(*n*) Determine the maximum value of $(I_{AM} + I_{AA})$ corresponding to $I_A'{}_{(max)}$.

$$(I_{AM} + I_{AA})_{(max)} = I_{K(max)} - I_A'{}_{(max)} \quad (8.10a)$$

(*o*) Determine the minimum value of $(I_{AM} + I_{AA})$ corresponding to $I_A'{}_{(min)}$

$$(I_{AM} + I_{AA})_{(min)} = I_{K(min)} - I_A'{}_{(min)} \quad (8.10b)$$

(*p*) Determine the approximate maximum value of the tolerance, w_M, on R_{AM} and $(R_{AA} + R_{AB})$.

$$w_{M(max)} \approx \frac{1}{2} \cdot \left[\frac{(I_{AM} + I_{AA})_{(max)}}{(I_{AM} + I_{AA})_{(min)}} \times \left\{ \frac{(1 - r)V_B - V_{MM} - I_K R_K}{(1 + r)V_B - V_{MM} - I_K R_K} \right\} - 1 \right] \quad (8.13)$$

(*q*) Determine R_{AM}.

$$R_{AM} = \frac{V_B - V_{MM} - I_K R_K}{\frac{1}{2}(I_{AM(max)} + I_{AM(min)})} \quad (8.7c)$$

(*r*) Determine maximum value, $I_{AM}'{}_{(max)}$, of I_{AM} corresponding to $I_A'{}_{(max)}$.

$$I_{AM}'{}_{(max)} = \frac{(1 + r)V_B - V_{MM} - I_K R_K}{(1 - w_M)R_{AM}} \quad (8.7d)$$

(*s*) Determine minimum value, $I_{AM}'{}_{(min)}$, of I_{AM} corresponding to $I_A'{}_{(min)}$.

$$I_{AM}'{}_{(min)} = \frac{(1 - r)V_B - V_{MM} - I_K R_K}{(1 + w_M)R_{AM}} \quad (8.7e)$$

(*t*) Using Equation (8.15a) (derived from Equations (8.8) and (8.10)), determine the maximum value of $(R_{AA} + R_{AB})$.

$$(1 + w_{AA})(R_{AA} + R_{AB})_{(max)} = \frac{(1 - r)V_B - V_{AA} - I_K R_K}{I_{K(min)} - I_A'{}_{(min)} - I_{AM}'{}_{(min)}} \quad (8.15a)$$

(*u*) Determine minimum value of $(R_{AA} + R_{AB})$.

$$(1 - w_{AA})(R_{AA} + R_{AB})_{(min)} = \frac{(1 + r)V_B - V_{AA} - I_K R_K}{I_{K(max)} - I_A'{}_{(max)} - I_{AM}'{}_{(max)}} \quad (8.15b)$$

(*v*) Using Equation (8.16) (derived from Equations (8.9) and (8.10)), determine the approximate value of R_{AA}.

$$R_{AA} \approx \frac{40}{\frac{1}{2}(I_{K(max)} + I_{K(min)}) - \frac{1}{2}(I_A'{}_{(max)} + I_A'{}_{(min)} + I_{AM}'{}_{(max)} + I_{AM}'{}_{(min)})} \quad (8.16)$$

(*w*) Choose R_{AA}, R_{AB}, and their tolerances to satisfy the results of steps (*t*), (*u*), and (*v*). R_{AA} should be made larger than the calculated value, rather than smaller.

EXAMPLE 8.2 Display Tube Directly Driven by Auxiliary-anode Stepping Tube

Design a circuit according to Fig. 8.14 to drive a GR10K display tube from an auxiliary-anode stepping tube, Type GSA10G.

(*a*) $I_{K(max)} = 3{\cdot}0$ mA $\qquad I_{AM(max)} = 0{\cdot}9$ mA

$I_{K(min)} = 2{\cdot}3$ mA $\qquad I_{AM(min)} = 0{\cdot}5$ mA

$V_{MM} = 240$ V $\qquad R_K = 3{\cdot}3$ kΩ

$V_{MA} = 225$ V

(*b*) $I_{A(max)} = 1{\cdot}8$ mA

$I_{A(min)} = 1{\cdot}0$ mA

$V_M = 140$ V

(*c*)
$$\frac{\delta I_K}{I_K} = \frac{3{\cdot}0 - 2{\cdot}3}{\frac{1}{2}(3{\cdot}0 + 2{\cdot}3)} = \mathbf{0{\cdot}264} = A$$

(*d*) Put $r \approx \frac{0{\cdot}264}{10} = \mathbf{0{\cdot}025}$, say (i.e. tolerance on V_B is $\pm 2\frac{1}{2}\%$). Then,

$$V_{B(nom)\,(min)} = \frac{0{\cdot}264}{0{\cdot}264 - 2 \times 0{\cdot}025} \cdot [225 + 140 + \tfrac{1}{2}(3{\cdot}0 + 2{\cdot}3) \times 3{\cdot}3]$$
$$= \mathbf{460\ V}$$

(*e*) Put $V_B = \mathbf{475\ V} \pm 2\frac{1}{2}\%$

(*f*) $$R_A \approx \frac{(475 - 225 - 140 - 9)}{\frac{1}{2}(1\cdot8 + 1\cdot0)} = 72\ \text{k}\Omega, \text{ say } \mathbf{68\ k\Omega}$$

(*g*) $$w_{A(max)} = \frac{0\cdot025 \times 475}{475 - 225 - 140} = \mathbf{0\cdot108}$$

(*h*) Put $w_A = 0\cdot10$ ($w_A < 0\cdot108$).

(*j*) $$I_A'_{(max)} = \frac{1\cdot025 \times 475 - 140 - 9}{0\cdot9 \times 68} = \mathbf{1\cdot85\ mA}$$

(*k*) $I_A'_{(max)} = 1\cdot85$ mA $\not< I_{A(max)}$. Hence reduce w_A to $w_A = 0\cdot05$ (i.e. $R_A = 68\ \text{k}\Omega \pm 5\%$).

(*j'*) $$I_A'_{(max)} = \frac{1\cdot025 \times 475 - 225 - 140 - 9}{0\cdot95 \times 68} = \mathbf{1\cdot75\ mA}$$

(*k'*) $$I_A'_{(max)} = 1\cdot75\ \text{mA} < I_{A(max)}$$

(*l*) $$I_A'_{(min)} = \frac{0\cdot975 \times 475 - 225 - 140 - 9}{1\cdot05 \times 68} = \mathbf{1\cdot26\ mA}$$

(*m*) $$I_A'_{(min)} = 1\cdot26\ \text{mA} > I_{A(min)}$$

(*n*) $$(I_{AM} + I_{AA})_{(max)} = 3\cdot0 - 1\cdot75 = \mathbf{1\cdot25\ mA}$$

(*o*) $$(I_{AM} + I_{AA})_{(min)} = 2\cdot3 - 1\cdot26 = \mathbf{1\cdot04\ mA}$$

(*p*) $$w_{M(max)} = \frac{1}{2} \cdot \left[\frac{1\cdot25}{1\cdot04} \cdot \left\{\frac{0\cdot975 \times 475 - 240 - 9}{1\cdot025 \times 475 - 240 - 9}\right\} - 1\right] = \mathbf{0\cdot0425}$$

As this value is slightly conservative, tentatively put $w_M = \mathbf{0\cdot05}$

(*q*) $$R_{AM} = \frac{475 - 240 - 9}{\frac{1}{2}(0\cdot9 + 0\cdot5)} = 320\ \text{k}\Omega, \text{ say } \mathbf{330\ k\Omega} \pm 5\%$$

(*r*) $$I_{AM}'_{(max)} = \frac{1\cdot025 \times 475 - 9}{0\cdot95 \times 330} = \mathbf{0\cdot76\ mA}$$

(*s*) $$I_{AM}'_{(min)} = \frac{0\cdot975 \times 475 - 240 - 9}{1\cdot05 \times 330} = \mathbf{0\cdot62\ mA}$$

(*t*) $$(1 + w_{AA})(R_{AA} + R_{AB})_{(max)} = \frac{0\cdot975 \times 475 - 225 - 9}{2\cdot3 - 1\cdot26 - 0\cdot62} = \mathbf{512\ k\Omega}$$

(*u*) $$(1 - w_{AA})(R_{AA} + R_{AB})_{(min)} = \frac{1\cdot025 \times 475 - 225 - 9}{3\cdot0 - 1\cdot75 - 0\cdot76} = \mathbf{485\ k\Omega}$$

(*v*) $$R_{AA} \approx \frac{40}{\frac{1}{2}(3\cdot0 + 2\cdot3) - \frac{1}{2}(1\cdot75 + 1\cdot26 + 0\cdot76 + 0\cdot62)} = \mathbf{88\ k\Omega}$$

(*w*) Put
$$R_{AA} = 110\ \text{k}\Omega \pm 5\%$$
$$R_{AB} = 390\ \text{k}\Omega \pm 2\%$$

Solution

In the circuit of Fig. 8.14,

V_1 = GR10K	$R_A = 63\ \text{k}\Omega \pm 5\%$
V_2 = GSA10G	$R_{AM} = 330\ \text{k}\Omega \pm 5\%$
$V_B = 475\ \text{V} \pm 12\ \text{V}$	$R_{AA} = 110\ \text{k}\Omega \pm 5\%$
$R_K = 3\cdot3\ \text{k}\Omega \pm 5\%$	$R_{AB} = 390\ \text{k}\Omega \pm 2\%$

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TABLE I

Voltage Stabilizing and Reference Tubes

V_M (V)	$I_{k\ min}$ (mA)	$I_{k\ max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
>50	—	0·3	75	WE	NE2L	General Elect.	—	—
50	0·1	0·5	90	WE	G50/1G‡	S.T. and C.	2208	—
54	—	0·1	70	WE	NE80	General Elect.	—	—
54	0·3	3·0	90	WE	G50/2G	S.T. and C.	2208	—
55	—	0·3	75	WE	NE68	General Elect.	—	—
55	—	0·3	75	WE	NE68A	General Elect.	—	—
>55	—	0·3	75	WE	NE2AS	General Elect.	—	—
>55	—	0·4	75	WE	NE75	General Elect.	—	—
55	—	0·4	72	WE	NE76	General Elect.	—	—
55	—	0·3	72	WE	NE81	General Elect.	—	—
55	—	1·5	72	WE	NE86	General Elect.	—	—
55	2·0	30	120	B4	G120/1B‡	S.T. and C.	548	—
55	2·0	30	90	B7G	G55/1K	S.T. and C.	5298	—
57	0·2	0·8	75	WE	VR57	Victoreen	—	—
59	—	0·3	75	WE	NE25	General Elect.	—	—
59	—	0·3	75	WE	NE23	General Elect.	—	—
60	—	1·0	80	WE	—	Hivac	8279	—
60	—	1·0	80	WE	NT2	Hivac	2213	—
60	—	1·0	160	WE	XC25	Hivac	—	—
60	—	2·0	80	WE	XC15	Hivac	5088	—
60	—	2·0	80	S.B.C.	XC16	Hivac	651	991
62	—	1·0	175	WE	XC26	Hivac	—	—
65	—	10	80	WE	NE83	General Elect.	—	—
70	—	0·5	135	WE	NE96	General Elect.	—	—
70	—	0·5	125	WE	NE97	General Elect.	—	—
70	—	0·75	163	WE	XC17‡	Hivac	5078	—

V_M (V)	$I_{k\ min}$ (mA)	$I_{k\ max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
70	—	0·75	145	WE	XC20‡	Hivac	—	—
75	—	0·75	157	WE	XC14‡	Hivac	—	—
75	2·0	20	110 (160)	B7G	QS75/20	Eng. Elect., G.E.C.	284	—
75	2·0	22	110	B7G	GTR75M	Ericsson	284	—
75	2·0	22	110	B7G	75B1‡	Mullard, Philips	284	—
75	5·0	30	115 (145)	B7G	OC2	Eng. Elect.	—	OC2
75	5·0	40	105	B8-0	VR75/30	S.T. and C.	3798	OA3
75	5·0	40	105 (160)	B8-0	OA3	Eng. Elect.	3798	OA3
75	5·0	40	105	B8-0	QS75/40	G.E.C.	3798	OA3
75	5·0	40	—	B8-0	OA3	Tung-sol	—	OA3
75	5·0	40	105	B8-0	QS1205	G.E.C.	3798	—
75	5·0	60	117	B8G	QS75/60	Eng. Elect., G.E.C.	434	—
75	2·0	60	115	B7G	SR51	Cerberus	—	OC2
75	2·0	60	115	B7G	GD75P	Ericsson	—	OC2
75	2·0	60	115	B7G	75C1	Mullard, Philips	—	—
75	2·0	60	110	B7G	M8225	Mullard, Philips	4080	—
75	5·0	60	115	WE	G75/3G	S.T. and C.	4030	—
81	2·0	4·0	120	WE	ZZ1000	Mullard, Philips	—	—
83	0·050	0·250	130	WE	GTR83X	Ericsson	—	—
83	0·050	0·250	95 (135)	WE	GTR83W	Ericsson	—	—
83	1·5	3·5	115	B7G	SR54	Cerberus	—	5651
83	1·0	5·0	125	B7G	QS83/3‡	Eng. Elect.	449	5651
83	3·5	6·0	130	B7G	SR52	Cerberus	—	—
83	3·5	6·0	130	B7G	83A1	Mullard, Philips	—	7980
83	3·5	6·0	130	B7G	GD83M	Ericsson	—	—
84	0·5	2·5	90	WE	SR42	Cerberus	—	—
84	0·5	5·0	115	WE	SR6	Cerberus	—	—
85	0·05	0·25	95	WE	SR41	Cerberus	—	—
85	—	1·0	— (104)	WE	XC12T	Hivac	5384	—
85	—	1·0	104	WE	XC12	Hivac	6004	—

* Where two figures are given, that in brackets relates to operation in dark.

† WE = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE I—*continued*

V_M (V)	$I_{k\ min}$ (mA)	$I_{k\ max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
85	0·5	3·5	125	WE	M8190	Mullard, Philips	4066	5783WA
85	0·5	5·0	110	WE	SR4	Cerberus	—	—
85	0·5	5·0	125	WE	GD85/WR	Ericsson	—	—
85	1·0	8·0	125	B8G	85A1‡	Mullard, Philips	431	OE3
85	1·0	10	125	B7G	SR5	Cerberus	—	—
85	1·0	10	115	B7G	SR53	Cerberus	—	OG3
85	1·0	10	115	B7G	QS83/3	G.E.C.	449	5651
85	1·0	10	115 (160)	B7G	QS1209	Eng. Elect., G.E.C.	449	5651
85	1·0	10	115	B7G	GD85M/S	Ericsson	449	OG3
85	1·0	10	115	B7G	85A2	Mullard, Philips	449	OG3
85	1·0	10	115	B7G	OG3	Tung-sol	—	OG3
85	1·0	10	115	B7G	GD85P/RS	Ericsson	4048	—
85	1·0	10	— (115)	B7G	QS1212	Eng. Elect., G.E.C.	4048	5651WA
85	1·0	10	115	B7G	M8098	Mullard, Philips	4048	—
85	1·0	10	— (115)	WE	QS1213	Eng. Elect., G.E.C.	4054	—
86	0·05	1·0	125	WE	GD86W/S‡	Ericsson	2321	—
86	1·5	3·5	—	B7G	5651	Mullard, Philips	—	5651
86	1·5	3·5	—	B7G	5651	Tung-sol	—	5651
86	1·5	3·5	—	B7G	5651WA	Tung-sol	—	5651WA
86	1·5	3·5	—	WE	5783	Tung-sol	—	5783
86	1·5	3·5	—	WE	5783WA	Tung-sol	—	5783WA
86	1·5	3·5	—	WE	5783WB	Tung-sol	—	5783WB
87	1·5	3·5	115	B7G	GD87M	Ericsson	2573	5651
87	2·0	80	130	B9A	ES1	Elesta	—	—
88	2·0	80	135	B9A	SR2A	Cerberus	—	—
90	0·1	5·0	175	WE	VX62	Victoreen	—	—
90	5·0	25	130	WE	5644	Mullard, Philips	3987	5644
90	1·0	40	— (115)	B7G	QS1215	Eng. Elect., G.E.C.	5173	—

V_M (V)	$I_{k\ min}$ (mA)	$I_{k\ max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
90	1·0	40	115	B7G	GD90M	Ericsson	—	—
90	1·0	40	115	B7G	90C1	Mullard, Philips	5173	—
90	1·0	40	115	B7G	M8206	Mullard, Philips	—	—
92	1·0	10	140	B4	QS92/10	Eng. Elect., G.E.C.	188/1070	—
95	2·0	10	110 (150)	B7G	QS95/10	Eng. Elect., G.E.C.	286	—
95	2·0	10	110 (150)	B7G	GTR95M/S	Ericsson	286	—
95	2·0	10	110 (150)	B7G	95A1‡	Mullard, Philips	286	—
97	0·5	8·0	135	WE	ES11	Elesta	—	—
98	5·0	25	—	WE	5787WA	Tung-sol	—	5787WA
100	1·0	8·0	140	B4	ST11	G.E.C.	—	—
100	1·0	8·0	140	B4	7475	Mullard, Philips	188/1070	—
100	1·0	10	140	B4	GD100A/GD100B‡	Ericsson	188/1070	—
100	10	40	130	Side contact	4687‡	Mullard, Philips	—	—
103	2·0	80	140	B9A	ES2	Elesta	—	—
104	0·5	5·0	140	WE	SR7	Cerberus	—	—
105	5·0	40	—	B8-0	OC3	Tung-sol	—	OC3
105	5·0	40	133	B8-0	G105/1D‡	S.T. and C.	686	OC3
105	5·0	40	—	B8-0	OC3W	Tung-sol	—	OC3W
107	2·0	80	155	B9A	SR3A	Cerberus	—	—
108	2·0	15	133	WE	QS1202	Eng. Elect., G.E.C.	4052	—
108	5·0	30	127	B7G	SR55	Cerberus	—	OB2
108	5·0	30	127	B7G	OB2	Eng. Elect., S.T. and C.	1833	OB2
108	5·0	30	127	B7G	GD108M	Ericsson	1833	OB2
108	5·0	30	133	B7G	QS1208	G.E.C.	1833	OB2
108	5·0	30	133	B7G	108C1	Mullard, Philips	1833	OB2
108	5·0	30	133	B7G	G108/1K	S.T. and C.	1833	OB2
108	5·0	30	—	B7G	OB2	Tung-sol	—	OB2
108	5·0	30	133	B7G	OB2WA	Eng. Elect.	4028	OB2WA

* Where two figures are given, that in brackets relates to operation in dark.
† WE = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE I—*continued*

V_M (V)	$I_{k\,min}$ (mA)	$I_{k\,max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
108	5·0	30	133	B7G	QS1211	G.E.C.	4028	OB2WA
108	5·0	30	133	B7G	M8224	Mullard, Philips	4028	OB2WA
108	5·0	30	—	B7G	OB2WA	Tung-sol	—	OB2WA
108	5·0	30	—	B7G	6627/OB2WA	Tung-sol	—	6627/OB2WA
108	5·0	30	—	B7G	6074	Tung-sol	—	6074
108	5·0	40	133	B8-0	OC3	Eng. Elect.	686	OC3
108	5·0	40	133	B8-0	QS1206	G.E.C.	686	OC3
108	5·0	40	127	B8-0	VR105/30	S.T. and C.	686	OC3
108	5·0	45	120 (150)	B8G	QS108/45	Eng. Elect., G.E.C.	422	—
120	10	75	135 (190)	B4	GTR120A/S‡	Ericsson	45	—
120	10	75	180	B4	GD120A/S‡	Ericsson	1110/1731	—
125	5·0	75	123 (140)	B4	S130P	G.E.C.	—	—
145	0·5	2·0	175	WE	SR43	Cerberus	—	—
150	0·500	2·0	170 (210)	WE	GTR150W	Ericsson	—	—
150	0·1	5·0	225	WE	VX64	Victoreen	—	—
150	5·0	15	180	B7G	QS1200	Eng. Elect., G.E.C.	2225	—
150	5·0	15	180	B7G	GD150P/S	Ericsson	2225	—
150	5·0	15	180	B7G	15OB2	Mullard, Philips	2225	6354
150	5·0	15	180	WE	QS1203	G.E.C.	4053	—
150	2·0	15	180 (225)	WE	QS1203	Eng. Elect.	4053	—
150	5·0	15	180	B7G	GD150PR/S	Ericsson	4104	—
150	5·0	15	180	B7G	SR57	Cerberus	—	—
150	5·0	15	180	B7G	M8163	Mullard, Philips	4104	—
150	2·0	15	170 (250)	B7G	QS150/15	G.E.C.	287	—
150	2·0	15	170 (240)	B7G	QS150/15	Eng. Elect.	287	—
150	2·0	20	170 (240)	B7G	GTR150M/S	Ericsson	287	—
150	2·0	20	170 (240)	B7G	150B3‡	Mullard, Philips	287	—
150	5·0	25	—	WE	6542	Tung-sol	—	6542

V_M (V)	$I_{k\,min}$ (mA)	$I_{k\,max}$ (mA)	V_{IG}* (V)	Base†	Type No.	Manufacturer	CV No.	American equivalent
150	5·0	30	180	B7G	SR56	Cerberus	—	OA2
150	5·0	30	185 (225)	B7G	OA2	Eng. Elect., S.T. and C.	1832	OA2
150	5·0	30	180	B7G	GD150M/S	Ericsson	1832	OA2
150	5·0	30	185	B7G	QS1207	G.E.C.	1832	OA2
150	5·0	30	185	B7G	150C2	Mullard, Philips	1832	OA2
150	5·0	30	185	B7G	G150/4K‡	S.T. and C.	1832	OA2
150	5·0	30	—	B7G	OA2	Tung-sol	—	OA2
150	5·0	30	165 (225)	B7G	OA2WA	Eng. Elect.	4020	OA2WA
150	5·0	30	165	B7G	QS1210	G.E.C.	4020	OA2WA
150	5·0	30	165	B7G	M8223	Mullard, Philips	4020/4100	OA2WA
150	5·0	30	185	B7G	150C4	Mullard, Philips	—	—
150	5·0	40	180 (225)	B8-0	OD3	Eng. Elect.	216	OD3
150	5·0	40	180	B8-0	GD150A/S	Ericsson	216	OD3
150	5·0	40	180	B8-0	QS150/40	G.E.C.	216	OD3
150	5·0	40	180	B8-0	150C3‡	Mullard, Philips	216	OD3
150	5·0	40	185	B8-0	G150/3D‡	S.T. and C.	216	OD3
150	5·0	40	180	B8-0	VR150/30	S.T. and C.	216	OD3
150	5·0	40	—	B8-0	OD3	Tung-sol	—	OD3
150	5·0	40	—	B8-0	OD3W	Tung-sol	—	OD3W
150	5·0	45	170 (200)	B8G	QS150/45	Eng. Elect., G.E.C.	395	—
150	5·0	45	180 (200)	B8G	G180/2M	S.T. and C.	395	—
150	5·0	45	180 (200)	WE	G180/2G	—	—	—
155	0·075	0·300	—	WE	7099	Tung-sol	—	7099
280	5·0	60/60/40/35	420	B5	STV280/40‡	G.E.C.	1068	—
280	10	100/90/70/70	420	B5	STV280/80‡	G.E.C.	1069	—
306	2·0	4·0	400	B7G	G400/1K	S.T. and C.	2194	—
306	2·0	4·0	400	WE	G400/2G	S.T. and C.	4047	—

* Where two figures are given, that in brackets relates to operation in dark.
† WE = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE II

Corona Stabilizers

V_M* (V)	$I_{k\,min}$ (mA)	$I_{k\,max}$ (mA)	V_{IG} (V)	Base†	Type No.	Manufacturer	CV No.
340	0·003	0·200		W.E.	GD340W	Ericsson	
350	0·001	0·100	400	W.E.	SC3/350	G.E.C.	
350	0·003	0·200		W.E.	GD350X	Ericsson	
350	0·003	0·200		W.E.	GD350Y	Ericsson	
350	0·001	0·325	400	B7G	SC1/350	G.E.C.	2456
400	0·001	0·100	450	W.E.	SC3/400	G.E.C.	
400	0·010	0·200		W.E.	GV4S-400	Victoreen	
400	0·005	0·300		W.E.	GV3B-400	Victoreen	
400	0·002	0·300		W.E.	GV3S-400	Victoreen	
400	0·001	0·350	450	B7G	SC1/400	G.E.C.	2487
400	0·010	0·400		B7G	GV5B-400	Victoreen	
400	0·050	1·000		B9A	GV6A-400	Victoreen	
450	0·100	1·500		B8-0	GV9A-450	Victoreen	
500		0·055	550	W.E.	G500/1G	S.T. and C.	
500	0·005	0·300		W.E.	GV3B-500	Victoreen	
600	0·003	0·100	650	W.E.	SC3/600	G.E.C.	
600	0·005	0·300		W.E.	GV3B-600	Victoreen	
600	0·003	0·300		W.E.	GV3S-600	Victoreen	
600	0·010	0·300		W.E.	GV4S-600	Victoreen	
600	0·004	0·400	650	B7G	SC1/600	G.E.C.	2458
600	0·050	0·750		B9A	GV6A-600	Victoreen	
600	0·100	1·500		B8-0	GV9A-600	Victoreen	
700	0·005	0·300		W.E.	GVB-700	Victoreen	
700	0·010	0·450		B7G	GV5B-700	Victoreen	

800	0·007	0·100	850	W.E.	SC3/800	G.E.C.	
800	0·010	0·375		W.E.	GV4S-800	Victoreen	
800	0·005	0·400		W.E.	GV3B-800	Victoreen	
800	0·003	0·450		W.E.	GV3S-800	Victoreen	
800	0·011	0·475	850	B7G	SC1/800	G.E.C.	2459
800	0·050	1·000		B9A	GV6A-800	Victoreen	
900	0·010	0·500		W.E.	GV3A-900	Victoreen	
900	0·010	0·550		B7G	GV5B-900	Victoreen	
1,000		0·100	1,100	W.E.	G1000/1G	S.T. and C.	
1,000	0·009	0·125	1,050	W.E.	SC3/1000	G.E.C.	
1,000	0·010	0·500		W.E.	GV3A-1000	Victoreen	
1,000	0·005	0·500		W.E.	GV3S-1000	Victoreen	
1,000	0·010	0·500		W.E.	GV4S-1000	Victoreen	
1,000	0·014	0·550	1,050	B7G	SC1/1000	G.E.C.	2460
1,000	0·010	0·650		B7G	GV5A-1000	Victoreen	
1,000	0·050	1·000		B9A	GV6A-1000	Victoreen	
1,000	0·100	1·750		B8-0	GV9A-1000	Victoreen	
1,200	0·011	0·125	1,250	W.E.	SC3/1200	G.E.C.	
1,200	0·015	0·600		W.E.	GV3A-1200	Victoreen	
1,200	0·010	0·600		W.E.	GV3S-1200	Victoreen	
1,200	0·015	0·600		W.E.	GV4S-1200	Victoreen	
1,200	0·016	0·625	1,250	B7G	SC1/1200	G.E.C.	2461
1,200	0·015	0·700		B7G	GV5A-1200	Victoreen	
1,200	0·020	0·700		B7G	QS1218/1200	Eng. Elect.	

* As the stabilizing voltage, V_M, is a function of gas pressure, manufacturers can provide tubes to operate at any required intermediate voltage.

† W.E. = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

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TABLE II—*continued*

V_M* (V)	$I_{k\ min}$ (mA)	$I_{k\ max}$ (mA)	V_{IG} (V)	Base†	Type No.	Manufacturer	CV No.
1,200	0·020	0·700		W.E.	QS1221/1200	Eng. Elect.	
1,200	0·050	1·200		B9A	GV6A-1200	Victoreen	
1,400	0·011	0·125	1,450	W.E.	SC3/1400	G.E.C.	
1,400	0·016	0·700	1,450	B7G	SC1/1400	G.E.C.	2462
1,400	0·015	0·700		W.E.	GV3A-1400	Victoreen	
1,400	0·020	0·700		B7G	QS1218/1400	Eng. Elect.	
1,400	0·020	0·700		W.E.	QS1221/1400	Eng. Elect.	
1,400	0·050	1·600		B9A	GV6A-1400	Victoreen	
1,500		0·100	1,650	W.E.	G1500/1G	S.T. and C.	
1,500	0·015	0·650		W.E.	GV3S-1500	Victoreen	
1,500	0·100	2·000		B8-0	GV9A-1500	Victoreen	
1,600	0·011	0·150	1,650	W.E.	SC3/1600	G.E.C.	
1,600	0·016	0·750	1,650	B7G	SC1/1600	G.E.C.	6065
1,600	0·020	0·800		W.E.	GV4S-1600	Victoreen	
1,600	0·020	0·800		B7G	GV5A-1600	Victoreen	
1,600	0·020	0·850		B7G	QS1218/1600	Eng. Elect.	
1,600	0·020	0·850		W.E.	QS1221/1600	Eng. Elect.	
1,600	0·050	1·800		B9A	GV6A-1600	Victoreen	
1,800	0·011	0·150	1,850	W.E.	SC3/1800	G.E.C.	
1,800	0·020	0·550		W.E.	GV3A-1800	Victoreen	
1,800	0·020	0·550		W.E.	GV3S-1800	Victoreen	
1,800	0·016	0·800	1,850	B7G	SC1/1800	G.E.C.	6066
1,800	0·020	0·900		B7G	QS1218/1800	Eng. Elect.	
1,800	0·020	0·900		W.E.	QS1221/1800	Eng. Elect.	
1,800	0·050	2·000		B9A	GV6A-1800	Victoreen	

1,850		0·100	1,980	W.E.	G1800/1G	S.T. and C.	
2,000	0·011	0·150	2,050	W.E.	SC3/2000	G.E.C.	
2,000	0·020	0·500		W.E.	GV3A-2000	Victoreen	
2,000	0·020	0·500		W.E.	GV3S-2000	Victoreen	
2,000	0·016	0·850	2,050	B7G	SC1/2000	G.E.C.	6067
2,000	0·020	1·000		W.E.	GV4S-2000	Victoreen	
2,000	0·020	1·000		B7G	GV5A-2000	Victoreen	
2,000	0·020	1·200		B7G	QS1218/2000	Eng. Elect.	
2,000	0·020	1·200		W.E.	QS1221/2000	Eng. Elect.	
2,000	0·050	2·000		B9A	GV6A-2000	Victoreen	
2,000	0·100	3·000		B8-0	GV9A-2000	Victoreen	
2,200	0·025	0·450		W.E.	GV3A-2200	Victoreen	
2,200	0·025	0·450		W.E.	GV3S-2200	Victoreen	
2,200	0·050	1·800		B9A	GV6A-2200	Victoreen	
2,400	0·025	0·410		W.E.	GV3S-2400	Victoreen	
2,400	0·025	1·000		W.E.	GV4S-2400	Victoreen	
2,400	0·025	1·000		B7G	GV5A-2400	Victoreen	
2,400	0·050	1·600		B9A	GV6A-2400	Victoreen	
2,500	0·025	0·400		W.E.	GV3A-2500	Victoreen	
2,500	0·025	1·000		B9A	SC2/2500	G.E.C.	
2,500	0·100	3·000		B8-0	GV9A-2500	Victoreen	
2,600	0·025	0·380		W.E.	GV3S-2600	Victoreen	
2,600	0·030	0·950		W.E.	GV4S-2600	Victoreen	
2,600	0·030	0·950		B7G	GV5A-2600	Victoreen	

* As the stabilizing voltage, V_M, is a function of gas pressure, manufacturers can provide tubes to operate at any required intermediate voltage.

† W.E. = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE II—*continued*

V_M* (V)	$I_{k\,min}$ (mA)	$I_{k\,max}$ (mA)	V_{IG} (V)	Base†	Type No.	Manufacturer	CV No.
2,600	0·050	1·500		B9A	GV6A-2600	Victoreen	
2,800	0·025	0·350		W.E.	GV3S-2800	Victoreen	
2,800	0·030	0·900		W.E.	GV4S-2800	Victoreen	
2,800	0·030	0·900		B7G	GV5A-2800	Victoreen	
3,000	0·025	0·330		W.E.	GV3S-3000	Victoreen	
3,000	0·030	0·850		W.E.	GV4S-3000	Victoreen	
3,000	0·030	0·850		B7G	GV5A-3000	Victoreen	
3,000	0·025	1·000		B9A	SC2/3000	G.E.C.	5844
3,000	0·050	1·300		B9A	GV6A-3000	Victoreen	
3,000	0·100	2·500		B8-0	GV9A-3000	Victoreen	
3,200	0·035	0·750		W.E.	GV4S-3200	Victoreen	
3,200	0·035	0·800		B7G	GV5A-3200	Victoreen	
3,500	0·035	0·700		B7G	GV5A-3500	Victoreen	
3,500	0·035	0·750		W.E.	GV4S-3500	Victoreen	
3,500	0·025	1·000		B9A	SC2/3500	G.E.C.	
4,000	0·050	0·600		B9A	GV6C-4000	Victoreen	
4,000	0·025	1·000		B9A	SC2/4000	G.E.C.	
5,000	0·050	0·500		B9A	GV6C-5000	Victoreen	
5,000	0·025	1·000		B9A	SC4/5000‡	G.E.C.	
5,000	0·050	1·000		End caps	SC5/5000	G.E.C.	
6,000	0·050	0·500		B9A	GV6C-6000	Victoreen	
6,000	0·025	1·000		B9A	SC4/6000‡	G.E.C.	
6,000	0·025	1·000		Special	M42C-6000	Victoreen	
6,000	0·050	1·000		End caps	SC5/6000	G.E.C.	
7,000	0·025	1·000		B9A	SC4/6800‡	G.E.C.	

7,000	0·050	1·000		End caps	SC5/6800	G.E.C.	
8,000	0·025	1·000		Special	M45C-8000	Victoreen	
10,000	0·025	1·000		Special	M45C-10,000	Victoreen	
10,000	0·050	1·000		Special	M108-10,000	Victoreen	
12,000	0·025	1·000		Special	M45C-12,000	Victoreen	
12,000	0·050	1·000		Special	M105-12,000	Victoreen	
12,000	0·050	1·000		Special	M108-12,000	Victoreen	
14,000	0·050	1·000		Special	M108-14,000	Victoreen	
16,000	0·050	1·000		Special	M105-16,000	Victoreen	
16,000	0·050	1·000		Special	M108-16,000	Victoreen	
18,000	0·050	1·000		Special	M105-18,000	Victoreen	
18,000	0·050	1·000		Special	M108-18,000	Victoreen	
20,000	0·050	1·000		Special	M105-20,000	Victoreen	
20,000	0·050	1·000		Special	M108-20,000	Victoreen	
20,000	0·050	1·000		Special	M126-20,000	Victoreen	
20,000	0·050	1·000		Special	M128-20,000	Victoreen	
22,000	0·050	1·000		Special	M126-22,000	Victoreen	
22,000	0·050	1·000		Special	M128-22,000	Victoreen	
24,000	0·050	1·000		Special	M126-24,000	Victoreen	
24,000	0·050	1·000		Special	M128-24,000	Victoreen	
25,000	0·050	1·000		Special	M126-25,000	Victoreen	
26,000	0·050	1·000		Special	M128-26,000	Victoreen	
27,000	0·050	1·000		Special	M128-27,000	Victoreen	

* As the stabilizing voltage, V_M, is a function of gas pressure, manufacturers can provide tubes to operate at any required intermediate voltage.

† W.E. = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE III

Relay (Trigger Glow) Tubes

$I_{k\,max}$ Av. (mA)	$I_{k\,max}$ Pk. (mA)	V_{IG} (V)	V_N (V)	V_T (V)	V_N (V)	Priming electrode	Remarks	Base†	Commercial code	Manufacturer	CV No.	American equivalent
0·5	—	190	85	115	—	—	—	W.E.	NE77	G.E.C.	—	—
0·5	—	210	70	69	55	—	—	W.E.	XC22‡	Hivac	—	—
1·0	—	200	77	75	—	—	—	W.E.	XC11‡	Hivac	—	—
1·0	—	210	73	68	55	—	—	W.E.	XC18	Hivac	2486	—
1·0	—	210	73	68	55	—	Twin trigger tube	W.E.	XC24‡	Hivac	—	—
1·5	—	235	70	85	57	—	—	W.E.	G1/236G‡	S.T. and C.	3524	—
1·5	—	200	68	75	59	—	—	W.E.	G1/237G	S.T. and C.	—	—
1·5	—	200	68	75	59	—	—	W.E.	G1/238G	S.T. and C.	—	—
2·0	—	310	120	145	—	k	—	W.E.	XC31	Hivac	—	—
2·5	10	275	110	146	—	—	—	B9A	Z800U‡	Mullard, Philips	2236	6538
2·5	10	170	105	*	—	—	45 $\mu\mu$C negative-going pulse on auxiliary cathode	B9A	Z801U‡	Mullard, Philips	2255	6539
3·5	50*	310	150	178	135	k	Over 20 μsec max	B7G	GTE175M	Ericsson	5348	—
3·5	50*	310	150	178	135	k	Over 20 μsec max; twin trigger tube	B7G	GPE175M	Ericsson	—	—
4·0	16	310	116	145	115	k	—	W.E.	Z700U	Mullard	5820	7710
4·0	16	310	116	145	115	k	—	W.E.	Z70U	Philips	5820	7710
4·0	16	310	116	145	115	k	Twin trigger tube	W.E.	Z700W	Mullard	—	7709
4·0	20	315	108	135	—	—	Pre-strike current ~10^{-12} A	B9A	GR19	Cerberus	—	—
5·0	—	275	107	118	95	a/k	Use and polarity of primer optional	W.E.	GTE120Y	Ericsson	—	—
5·0	>10	275	107	118	—	a	—	W.E.	GR43	Cerberus	—	—
6·0	20	330	115	130	—	a	Shielded-anode tube	B9A	GR18	Cerberus	—	—
7·0	9·0	165	60	80	64	—	Twin trigger speech tube	W.E.	Z701U	Mullard	—	7711
7·0	9·0	165	60	80	64	—	Twin trigger speech tube	W.E.	Z71U	Philips	—	7711
7·5	—	200	70	75	55	—	—	W.E.	XC13‡	Hivac	—	—
7·5	—	200	67	70	55	—	—	—	XC23	Hivac	5217	—
8·0	>100	290	110	145	—	a	Twin trigger tube	W.E.	GR21‡	Cerberus	—	—
9·0	—	310	117	131	63* 73*	—	At I_a = 0 mA; At I_a = 4·5 mA	W.E.	GTR120W	Ericsson	—	—
9·0	—	340	113	112	—	a and k*	Separate priming diode gap	B7G	GDT120M	Ericsson	—	—
10	30	200	80	75	—	—	—	B8-0	CC2R‡	Hivac	—	—

$I_{k\,max}$ Av. (mA)	$I_{k\,max}$ Pk. (mA)	V_{IG} (V)	V_N (V)	V_T (V)	V_N (V)	Priming electrode	Remarks	Base†	Commercial code	Manufacturer	CV No.	American equivalent
10	15	360	140	165	—	a and k*	High-speed, shielded-anode tube, separate priming diode gap	B7G	G1/371K	S.T. and C.	2224	—
10	>20	400	110	130	—	a	Twin trigger tube	W.E.	GR41	Cerberus	—	—
15	—	260	80	—	—	—	External touch-button	W.E.	GK11	Cerberus	—	—
15	—	400	115	130	—	a	Twin trigger tube	W.E.	ER32	Elesta	—	—
15	—	300	107	130	—	a	Twin trigger tube	W.E.	ER33	Elesta	—	—
25	—	180	72	80	—	—	External touch-button	B8-0	XC19	Hivac	—	—
25	100	200	85	105	—	—	—	B7G	5823	Eng. Elect.	—	5823
25	100	210	68	80	—	—	—	B7G	QT1250	Eng. Elect.	—	—
25	100	210	68	80	—	—	—	W.E.	QT1251	Eng. Elect.	—	—
25	60	400	112	127	—	a and k*	Separate priming diode gap	B9A	GDT120T	Ericsson	—	—
25	60*	290	105	132	—	a	1 A for 1 msec	B9A	GTE130T	Ericsson	2434	6779
25	200*	290	105	132	95	a	1 A for 1 msec	B9A	Z803U	Mullard, Philips	2434	6779
25	125	275 r.m.s.	112	−123	—	—	—	B9A	Z804U‡	Mullard, Philips	—	7713
25	100	225	70	80	—	—	—	B8-0	Z300T	Mullard, Philips	1992	1267
25	150	200	62	80	60	—	—	B7G	Z900T	Mullard, Philips	5122	5823
25	200	400	118	119	95	a	Shielded-anode tube	B9A	Z806W	Mullard, Philips	—	—
25	125	450	110	123	—	a	Shielded-anode tube	B9A	GR32	Cerberus	—	—
25	60	290	105	132	—	a	—	B9A	GR33	Cerberus	—	—
30	>2A	300	109	130	—	a	Twin trigger tube	B9A	GR20	Cerberus	—	—
30	50	150	68	70	60	—	—	B8-0	G150/2D	S.T. and C.	413	—
30	50	230	90	75	65	—	—	B8-0	G240/2D	S.T. and C.	2174	—
40	>2A	300	107	130	—	a	—	B9A	GR15	Cerberus	—	—
40	—	300	107	130	—	a	—	B9A	ER1	Elesta	—	—
40	—	300	107	130	—	a	Twin trigger tube	B9A	ER3	Elesta	—	—
40	>2A	370 (260 r.m.s.)	111	130 (92 r.m.s.)	—	a	For use on either d.c. or unrectified a.c. anode supply	B9A	GR16	Cerberus	—	—
40	—	400	111	130	—	a	—	B9A	ER2	Elesta	—	—
40	—	280 r.m.s.	111	100 r.m.s.	—	—	For unrectified a.c. anode supply	B9A	ER21A	Elesta	—	—
40	>2A	300 r.m.s.	113	−130	—	—	For unrectified a.c. anode supply. Strikes with anode +ve, trigger −ve	B9A	GR17	Cerberus	—	—
40	—	300 r.m.s.	112	−130	—	—	For unrectified a.c. anode supply. Strikes with anode +ve, trigger −ve	B9A	ER22	Elesta	—	—
40	>2A	400	111	130	—	a	—	B9A	GR31	Cerberus	—	—
40	—	450	115	−3	—	k	Glow thyratron	B9A	GT21	Cerberus	—	—
50‡	250*	390	105	123	—	a	Shielded-anode tube, 25 mA av. and 125 mA pt. for long-life applications	B9A	GPE120T	Ericsson	—	—

* See Remarks. † W.E. = Wire-ended or flying-lead connexion. ‡ Maintenance, obsolescent, or obsolete type.

TABLE IV

Stepping Tubes

$f_{(max)}$ (kc/s)	Index cathodes	$I_{a(min)}$ (mA)	$I_{a(max)}$ (mA)	V_M (V)	$V_{B(min)}$ (V)	Function	Type No.	Manu-facturer	CV No.	American equivalent
1	10	0·315	0·500	190	320	Reversible computer	GC10/2P*	Ericsson	—	—
1	10	3·0 nominal		95	150	Reversible selector with routing guides	GS10J	Ericsson	—	—
1	10	—	0·550	190	400	Non-reversible counter requiring no inter-stage amplifier. Single pulse	Z302C*	Mullard	—	—
4	10	0·250	0·550	191	350	Reversible counter	GC10B/S	Ericsson	2271	6482
4	10	0·250	0·550	191	350	Reversible counter	Z303C	Mullard	2271	6482
4	10	0·250	0·550	190	350	Reversible counter	GC10B/L	Ericsson	6044	—
4	10	0·250	0·550	191	350	Reversible computer	GC10/4B	Ericsson	1739	6802
4	10	0·250	0·550	190	350	Reversible computer	GC10/4B/L	Ericsson	6100	—
4	12	0·250	0·550	191	350	Reversible computer	GC12/4B	Ericsson	—	—
4	10	0·250	0·550	192	400	Reversible selector	GS10C/S	Ericsson	2325	6476
4	10	0·250	0·550	191	400	Reversible selector	Z502S	Mullard	2325	6476
4	12	0·190	0·350	191	400	Reversible selector	GS12D	Ericsson	—	—
4	12	0·250	0·550	192	350	Reversible selector	GS12C*	Ericsson	—	—
5	10	0·300	0·600	187	350	Reversible computer	6802	Raytheon	—	6802
5	10	0·250	0·525	195	375	Reversible selector	Z504S (ZM1070)	Mullard	—	—
5	10	0·250	0·525	195	375	Reversible selector	GS10M	Ericsson	—	—
5	10	0·300	0·600	187	350	Reversible selector	6476	Raytheon	—	6476
5	10	0·300	0·600	187	350	Reversible selector	6476A	Raytheon	—	6476A
5	10	0·300	0·600	190	350	Reversible selector	7978	Raytheon	—	7978
5	10	0·250	0·370	187	380	Reversible selector with routing guides	GS10H	Ericsson	—	—
10	10	1·500	2·000	210	480	Non-reversible selector, single pulse	GS10K*	Ericsson	—	—
10	10	0·700	0·900	208	440	Reversible selector	GS10E	Ericsson	—	—
10	10	0·700	0·900	210	400	Reversible selector with routing guides	GS10G*	Ericsson	—	—
10	10	*Main Anode* 0·500 / *Auxiliary Anode* 1·400	0·900 / 2·500	240 / 225	440	Reversible counter with auxiliary anodes and routing guides	GCA10G	Ericsson	—	—
10	10	*Main Anode* 0·500 / *Auxiliary Anodes* 1·400	0·900 / 2·500	240 / 225	440	Reversible selector with auxiliary anodes and routing guides	GSA10G	Ericsson	—	—
20	10	0·700	1·200	215	420	Non-reversible counter, single pulse	GC10D	Ericsson	5143	—
20	10	2·400	5·000	180	280	Non-reversible selector, single pulse	G10/241E	S.T. and C.	X2223	—
20	10	0·700	0·900	208	440	Reversible selector	GS10D	Ericsson	—	—
50	10	0·600	1·000	260	400	Reversible selector	Z505S (ZM1060)	Mullard	—	—
100	10					Non-reversible selector	EZ10	Elesta	—	—
100	10	0·600	0·800	235	400	Reversible computer	6909	Raytheon	—	6909
100	10	0·600	0·800	235	400	Reversible selector	6910	Raytheon	—	6910
100	10	0·600	0·800	235	400	Reversible selector	8262	Raytheon	—	8262
1,000	10	1·200	1·900	300	450	Non-reversible selector	EZ10B	Elesta	—	—
1,000	10	1·700	3·500	307	390	Selector. Reversible by switching connexions	ECT100	Elesta	—	—

* Maintenance, obsolescent or obsolete type.

TABLE V

Register and Display Tubes

No. of cathodes	V_M (V)	$I_{k(min)}$ (mA)	$I_{k(max)}$ (mA)	Pre-bias (V)	V_{IG} (V)	Display	Character height (mm)	Type No.	Manufacturer	CV No.	American equivalent
2	112	0·250	0·400	5	150	Off–on	—	TG121A	Fujitsu	—	—
1	75	—	5·0	20	140	Off–on	—	395A	Tung-sol	—	395A
1	—	—	12·0	—	180	Off–on	—	7400	Tung-sol	—	7400
1	110	1·0	8·0	5	180	Off–on	—	7401	Tung-sol	—	7401
1	—	—	—	—	—	Off–on	—	7813	Tung-sol	—	7813
10	108	0·050	0·250	24	129	Glow-position	—	GR10A	Ericsson	5291	—
10	108	0·050	0·250	24	129	Glow-position	—	Z503M	Mullard, Philips	5291	—
10	—	—	—	5	125 (90 a.c. rectified)	Clock-face 0–9, end view	—	Z550M	Mullard, Philips	—	—
10	—	0·250	0·400	—	150	Clock-face 0–9, end view	—	B9012	Burroughs	—	—
10	130	—	2·0	—	180	0–9, side view	14	XN1*	Hivac	—	—
10	140	1·25	2·0	60	180	0–9, side view	14	XN3	Hivac	—	—
10	—	1·25	2·5	—	200	0–9, side view	14	GN6	S.T. and C.	—	—
10	140	1·5	2·5	60	170	0–9, side view	13	ZM1080	Mullard, Philips	—	—
10	140	2·0	4·0	40	150	0–9, side view	20·5	GR10P	Ericsson	—	—
10	160	2·5	4·0	40	220	0–9, side view	15	GR10W	Ericsson	—	—
10	145	2·5	4·0	40	150	0–9, side view	30	GR10J	Ericsson	—	—
10	140	3·0	6·0	60	160	0–9, side view	31	Z522M (ZM1040)	Mullard, Philips	—	—
10	150	4·0	7·0	50	250	0–9, side view	50	B7037	Burroughs	—	—
10	180	5·5	9·0	—	220	0–9, side view	30	GR10G	Ericsson	—	—
10	135	5·0 nominal		—	170	0–9, side view	60	GR10N	Ericsson	—	—
12	140	1·25	2·0	60	180	0–11, side view	14	XN2	Hivac	—	—
10	102	0·7	1·2	50	170	0–9, end view	8	BD-200S	Burroughs†	—	7009
10	102	0·7	1·4	50	170	0–9, end view	8	B4081	Burroughs†	—	—
10	154	0·7	1·4	50	170	0–9, end view	8	B4032	Burroughs†	—	7977
10	102	0·7	1·4	50	120	0–9, end view	8	B4021	Burroughs†	—	—
10	—	0·7	1·4	—	170	0–9, end view, rectangular	8	B4991	Burroughs†	—	—
10	140	1·0	1·8	40	150	0–9, end view	19	GR10K	Ericsson	5842	—
10	140	1·0	2·5	40	150	0–9, end view	19	GR10H	Ericsson	—	—
10	140	1·0	2·5	60	170	0–9, end view	15·5	GR10M	Ericsson	—	—
10	140	1·0	2·5	60	170	0–9, end view	15·5	Z520M (ZM1020)	Mullard, Philips	5278	8037‡
10	—	1·5	3·0	—	170	0–9, end view	15·5	GN4	S.T. and C.	—	—
10	—	1·5	3·0	75	170	0–9, end view	15·5	GN3*	S.T. and C.	—	—
10	127	1·5	3·0	100	170	0–9, end view	15·5	Z510M*	Mullard, Philips	—	6844‡
10	135	1·5	3·0	50	170	0–9, end view	15·5	HB-106	Burroughs†	—	6844
10	135	1·5	3·0	50	170	0–9, end view	15·5	BD-302	Burroughs†	—	6844A
10	147	1·5	3·0	50	170	0–9, end view	15·5	B5031	Burroughs†	—	8037
10	147	1·5	3·0	50	170	0–9, end view	15·5	B5092	Burroughs†	—	8421
10	147	1·5	3·0	50	170	0–9, end view rectangular	15·5	B5991	Burroughs†	—	—
10	143	2·0	3·0	50	250	0–9, end view	20·5	BD-206	Burroughs†	—	7153
10	147	1·5	4·0	50	170	0–9, end view	20·5	B6033	Burroughs†	—	—
10	147	1·5	4·0	50	170	0–9, end view	20·5	B6091	Burroughs†	—	8423
10	135	2·5	5·0	—	200	0–9, end view for a.c. or d.c.	25·4	GN1*	S.T. and C.	—	—
10	—	2·5	nominal	—	200	0–9, end view	25·4	GN5	S.T. and C.	—	—
10	135	2·5	5·0	—	200	0–9, end view	25·4	GN2*	S.T. and C.	—	—
10	145	3·0	6·0	50	170	0–9, end view	25	B8091	Burroughs†	—	—
10	150	4·0	7·0	50	300	0–9, end view	50	B7094*	Burroughs†	—	—
2	140	1·25	2·0	60	180	½, side view	14	XN4	Hivac	—	—
4	170	4·3	7·0	40	200	¼, ½, ¾, 1, side view	30	CR4G	Ericsson	—	—
10	130	1·5	2·5	—	200	A to J, end view	19	GA1*	S.T. and C.	5237	—
10	130	1·5	2·5	—	200	K to T end view	19	GA2*	S.T. and C.	5238	—
10	147	1·5	3·0	50	170	A to K, not I, end view	15·5	B50322	Burroughs†	—	—
10	147	1·5	3·0	50	170	L to X, not O, Q, U, end view	15·5	B5035	Burroughs†	—	—
12	170 to 185	4·0	9·0	Switched	220	A to L, side view	30	GR12G	Ericsson	—	—
12	175	4·0	9·0	Switched	220	E, L to X, not P and Q, side view	30	GR12H	Ericsson	—	—
2	168	2·5	3·5 (−) 5·5 (+)	40	180	+, −, side view	18	GR2G	Ericsson	—	—
2	154	0·7	1·4	50	150	+, −, end view	8	B4031	Burroughs	—	—
2	130	0·75	1·5 (−) 2·0 (+)	40	150	+, −, end view	20	GR2H	Ericsson	—	—
2	140	1·0	2·0	—	200	+, −, end view	12	GS1	S.T. and C.	—	—
2	127	1·5	3·0	50	170	+, −, end view	15·5	B5016*	Burroughs†	—	—
2	147	1·5	3·0	50	170	+, −, end view	15·5	B50911	Burroughs†	—	—
2	147	1·5	3·0	50	170	+, −, end view rectangular	15·5	B5992	Burroughs†	—	—
7	140	1·0	2·5	60	160	+, −, V, A, Ω, %, ~, end view	15·5	Z521M (ZM1021)	Mullard	—	—
10	154	0·7	1·4	50	170	Two-character tube, *left*: K, M, μ, +, −, ~, *right*: C, S, V, Ω, end view	8	B50925	Burroughs†	—	—
13	120	3·0	7·0	50	170	Alpha-numeric 13-bar matrix	—	B5971	Burroughs	—	—

* Maintenance, obsolescent or obsolete type.
† In addition to those listed above, Burroughs have produced several hundred special character tubes providing various combinations of symbols.
‡ Mechanical dimensions differ.

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